Differential Detection for Differential orthogonal Space-Time Modulation with APSK signals

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CERTIFICATE

It is certified that the work contained in the thesis entitled "*Differential Detection for Differential orthogonal Space-Time Modulation with APSK signals* " by *Srikanth Katkam* has been carried out under my supervision and that this work has not been submitted elsewhere for a degree.

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Abstract

Differential space time modulation has been recently proposed for multiple antenna systems, when channel information is not available at receiver. A decision feedback based Amplitude Phase Shift Keying (APSK) modulation/demodulation has also been proposed in the literature, to improve the throughput of the system. It is well known that the performance of differential schemes is worse when compared to their non-differential counterparts. To improve the performance of the above system, we have proposed a modification (with theoretical proof) in the decoding technique of the Decision Feedback Differential Decoding for Differential Orthogonal Space Time modulation method with APSK signals.

Next we have studied the effect of replacing the Recursive Least Square (RLS) algorithm used in the decision feedback differential detection, with the simpler Least Mean Square (LMS) algorithm. The performance of both the systems are evaluated for different orders of the adaptive algorithm and under different channel conditions, i.e., slow fading channel (where the fading coefficients of the channel are constant for at least 2 blocks of data transmission) and fast fading channel (where the fading coefficients are varying for each block of the data transmission).

In all the above discussed systems, the decoder is designed on the assumption that the channel is flat fading. To enable the system to work in a frequency selective fading channel, a new layer with Orthogonal Frequency Division Multiplexing (OFDM) modulation has been added and the performance of the same is evaluated.

Dedicated to

To my mom, Ms. Kalavathi Katkam and my late father Shri. Seetha Ramulu Katkam.

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Chapter 1

Introduction

1.1 Evolution of Multiple Input Multiple Output systems

Research in wireless communications has been constantly driven by the user demand for better quality, high data rates, simpler and economic gadgets. The channel impairments, which pose considerable challenge to a system designer are fading and interference. Of these fading poses much serious problem, causing the received signal amplitude to fluctuate rapidly.

Fading is induced by multipath phenomenon, which causes the signal to combine destructively. Fading can be either frequency selective or flat, depending on the coherence bandwidth (bandwidth over which channel fading characteristics remain same) and the signal bandwidth. If the signal bandwidth is greater than coherence bandwidth the channel is said to be frequency selective, else it is called as flat.

Diversity in various forms can be exploited to combat the fading. Diversity adds calculated amount of redundancy in the transmitted signal. This redundancy can be in one or more forms, time, frequency and space. The reliability of the link as well as throughput can be improved by effectively utilizing these diversities.

Channel coding (block or convolution) exploits time diversity, examples include: several block codes, trellis codes etc. Systems with multi-carrier configuration exploit the frequency

diversity, examples include MC-CDMA systems, OFDM based systems etc. MIMO(Multiple Input Multiple Output) systems exploit the diversity in space, with multiple antennas at transmitter and receiver.

When the coherence time (a measure of the time selectivity of the channel, time during which channel fading characteristics remain same) of the channel is high, then it renders no temporal diversity. Also, when the channel is flat, i.e., when there is no frequency selectivity, it does not render frequency diversity. In such a situation, spatial diversity is of paramount importance, since it does not incur a penalty in transmission time and bandwidth (unlike time/frequency diversities).

Communication using multiple antennas at the transmitter and/or receiver has been demonstrated to result in dramatic increase in data rates. These systems are referred to as MIMO systems, and they offer spatial diversity. Recent information-theoretic results have shown the enormous capacities that can be realized with such systems. Thus, in the few years since their inception, they have attracted an enormous amount of interest. Although a lot of effort is still needed for research and standardization before they can become ubiquitous.

1.1.1 Capacity of MIMO channels

MIMO systems can show enormous improvement over Single Input Single Output(SISO) systems in capacity, if the spatial diversity is properly utilized. In this section we try to quantify the shannon's channel capacity for MIMO channels. We assume that channel is known to the receiver but not to the transmitter.

Let H be the channel matrix of dimension $n_t \times n_r$ where n_t is the number of transmit antennas and n_r be of that of receiver. S be the transmitted symbol matrix (designed according to the coding technique used), N be the noise vector of appropriate dimension. The matrix H is composed with $n_t * n_r$ complex random variables (fading coefficient).

Then the received signal vector say Y can be given as

$$Y = HS + N \tag{1.1}$$

The information theoretic point of view of above system model allows us to get the channel capacity (MIMO version of shannon's capacity theorem). The standard formula for

the shannon capacity [1] expressed in bps/Hz of SISO flat fading is:

$$C(h) = log_2(1 + \rho|h|^2) \tag{1.2}$$

where ρ is the SNR(Signal to Noise Ratio). It is evident that for high SNRs every 3 dB increase in ρ gives an extra one bps/Hz capacity.

Let S be transmitted vector with covariance $E(SS^H) = P$. Then the capacity of a fixed MIMO channel with Gaussian noise is given by [2] [3]:

$$C(H) = log_2(I + \frac{P}{\alpha^2 n_t} HH^H) \tag{1.3}$$

Where α is the noise power and I is the Identity matrix of appropriate dimensions. In order to make meaningful comparison with SISO case, we apply equation (1.3) to a special case, when $H = I_n$. Then, we get:

$$C(I_n) = n.log_2(1 + \frac{\rho}{n}) \longrightarrow \frac{\rho}{ln(2)} \tag{1.4}$$

as $n \longrightarrow \infty$, where $n = n_t = n_r$. It is evident from (1.4) that when we have orthogonal parallel channels the capacity grows linearly with SNR as compared to the logarithmic growth in SISO channel. It is also evident that for high SNRs a 3 dB increase in ρ gives another $n$bps/Hz capacity, compared to the unity increase in SISO case, i.e., capacity grows linearly with the number of antennas.

The spatial dimension offered by MIMO systems can be used to achieve significant improvements in *spectral efficiency* and *link reliability* by using *spatial multiplexing* and *space-time coding* respectively.

1.2 Spatial multiplexing

Spatial Multiplexing (SM) is a signalling technique to achieve a very high data rates using MIMO system. Bell labs LAyered Space Time(BLAST) is an extraordinarily bandwidth-efficient approach to wireless communication, which takes advantage of the spatial dimension by transmitting and detecting a number of independent co-channel data streams. BLAST system exploits(rather than mitigates) the multipath effects to achieve high spectral efficiencies.

There are two variants of BLAST: Vertical-BLAST(V-BLAST) [4] and Diagonal-BLAST(D-BLAST)[5].

D-BLAST utilizes multi-element antenna arrays at both transmitter and receiver and an elegant diagonally-layered coding structure in which code blocks are dispersed across diagonals in space-time. In an independent Rayleigh scattering environment, this processing structure leads to theoretical rate, which grows linearly with the number of antennas with these rates approaching 90% of the shannon capacity. However, the diagonal approach suffers from certain implementation complexities. V-BLAST is a simplified version of D-BLAST. It has been shown that it is possible to achieve spectral efficiencies of 20-40 bps/Hz at average SNRs ranging from 24-34 dB. The spatial dimension offered by MIMO system can also be used to improve the link reliability. Space time codes achieves this. Our focus in this work is on systems employing space-time codes.

1.3 Space Time Codes

Space Time Coding(STC) schemes allow for the adjusting and optimization of joint encoding across space and time in order to maximize the reliability of a wireless link.

Space Time Codes(STC) have been developed to harness the spatial diversity advantages offered by MIMO systems. There are two varieties of ST Codes available in the literature viz., Space-Time Block Codes (STBC) [6] and Space-Time Trellis Codes(STTC)[8].

Space-Time Block Codes: These codes are transmitted using an orthogonal block structure which enables simple decoding at the receiver.

Space-Time Trellis Codes: These are convolutional codes extended to the case of multiple transmit and receive antennas.

Although both share the common objective of achieving diversity gain, STBC results in simpler receivers whereas STTC achieve coding gain at the price of complexity of the receiver. We consider STBC in this work, owing to it's simplicity and wider appeal.

In the recent past, significant progress in code design of STC has been made. The primary focus was on the case when only the receiver knows the channel, which is the case for most practical systems. For this scenario, the first bandwidth efficient transmit diversity scheme

was proposed by Wittneben [13], and it includes the delay diversity scheme of Seshadri and Winters [14] as a special case. A transmit diversity scheme which achieves the same rate as that of antenna hopping diversity is constructed in [15]. Space-time trellis coding has been proposed in [8], which combines signal processing at the receiver with coding techniques appropriate to multiple transmit antennas, and it provides significant gain over [14] and [13]. The bandwidth efficiency for 2-3 antennas system is about 34 times that of current systems. The space-time codes presented in [8] provide the best possible tradeoff between constellation size, data rate, diversity advantage, and trellis complexity. When the number of transmit antennas is fixed, the decoding complexity of space-time trellis coding (measured by the number of trellis states at the decoder) increases exponentially as a function of both the diversity level and the transmission rate.
In addressing the issue of decoding complexity, Alamouti discovered a remarkable scheme for transmission using two transmit antennas [6]. This scheme supports a maximum likelihood detection scheme based only on linear processing at the receiver. Space-Time Block Coding introduced in [16] generalizes the transmission scheme discovered by Alamouti [6] to an arbitrary number of transmit antennas and is able to achieve the full diversity promised by the transmit and receive antennas. These codes retain the property of having a very simple maximum likelihood decoding algorithm based only on linear processing at the receiver [16]. For more details on transmit diversity when the receiver knows the channel, see [8] and [16] and the references therein.

Space-time codes are designed based on Rank and Determinant criteria [8]. Let S be the transmitted codeword and S_d the decoded word and $B = S - S_d$, is the error matrix. Both of the above depends on hermitian matrix $A = BB^H$.

Rank Criteria: In order to achieve the maximum diversity in a MIMO system the rank of the difference matrix B over all pairs of code matrices S and S_d should be maximized.

Determinant Criteria: If the code is designed to give a specific diversity gain, for a better coding gain, the minimum of the determinant of A taken over all pairs of distinct code matrices S and S_d must be maximized.

Both of these criteria apply to STBC as well as STTC. Appropriately designed STC code can give diversity gain as well as coding gain.

1.3.1 Space-Time Block Codes(STBC)

STBC have the advantage of simpler decoding compared to STTC. Orthogonal STBC (OSTBC) [16] [6] [17], which are special cases of STBC have been widely used because they result in simple receivers. OSTBC, however provide no coding gain (unlike trellis codes). OSTBC can be concatenated with powerful outer codes like turbo codes to achieve high coding gain. For complex constellations full rate codes exists only for $n_t = 2$ and for real constellations full rate codes exists only for $n_t = 2, 4$ and 8.

Alamouti [6] is a typical example of OSTBC. The Alamouti code considerably simplifies the receiver processing. In this work, we use Alamouti code because of it's simplicity. We consider a MIMO system with two transmit and one receive antenna.

Alamouti coding: we send two different symbols s_1 and s_2 on two different antennas, 1 and 2 simultaneously say at time t_1. Then in next symbol time ($t = t_1 + 1$) we send $-s_2*$ and s_1* from antennas 1 and 2 respectively. We assume that the channel remains constant for two symbol duration and is frequency flat. Let $h = [h_1\ h_2]$ be fading coefficients of 2 different channels, output y=$[y_1\ y_2]$ can be given as

$$\begin{bmatrix} y_1 & y_2 \end{bmatrix} = \begin{bmatrix} h_1 & h_2 \end{bmatrix} \begin{bmatrix} s_1 & -s_2* \\ s_2 & s_1* \end{bmatrix} + \begin{bmatrix} n_1 & n_2 \end{bmatrix} \tag{1.5}$$

The received vector can be rearranged to get

$$\begin{bmatrix} y_1 \\ y_2* \end{bmatrix} = \begin{bmatrix} h_1 & h_2 \\ h_2* & -h_1* \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2* \end{bmatrix} \tag{1.6}$$

Note that the transmission scheme orthogonalises the channel ($H_{eff}H_{eff}^H = ||h||_F^2 I_2$) irrespective of the channel realization (where $||h^2||_F$ denotes the Frobenius norm of h). The orthogonal channel is shown to have capacity maximizing property [7]. If $z = H_{eff}s$, we get

$$z = ||h||_F^2 I_2 s + \tilde{n} \tag{1.7}$$

Where I_2 is the identical matrix of dimension two and $\tilde{n} = H_{eff}^H n$. We notice from the above equation that the received SNR is increased by a factor $||h||_F^2$ and thus the Alamouti code is able to extract the full diversity present in the channel, even in the absence of channel knowledge at the transmitter.

1.3.2 Space-Time Trellis Codes

STTC [8] provide coding unlike OSTBC. The coding gain of the STTC depends on the number of states of the trellis. For higher coding gain a large number of states are required, and the complexity also increases exponentially. The STTC requires vector Viterbi decoding, which makes it computationally complex. In the entire work presented here, we stick to STBC only.

1.4 Challenges of Conventional Space Time Block Coded systems

Frequency offset between oscillators at transmitters and Fading coefficients of channel are the crucial parameters that decide the performance of a conventional STBC system. Most of the systems/designs proposed in the literature for STBC [6] [8] [13] [15] [16] assume that CSI (Channel State Information) is available at the receiver. They do coherent detection with assumption of perfect synchronization of oscillators. In practice it is left to the burden of signal processing engineers to estimate the channel fading coefficients and frequency offset amount. This problem has attracted considerable research attention and several algorithms have been proposed for channel and offset estimation [9] [10]. Most of these algorithms use training symbols, which consume considerable share of precious bandwidth.

There are some techniques in the literature which work fine with very less number of pilot symbols as in [18]. There are several blind (which need no pilot symbols at all) [19] [20] [21] and semi-blind(which need very few pilot signals) [22] [23] techniques in literature. But in general there are some limitations in these techniques that, these not only results in complex receivers but also requires channel to be constant for longer period of time. In [20] the estimation process is made completely blind, but receiver is extremely complicated. Also, results in unpleasantly high NMSE (Normalized Mean Square Error) in channel estimation at lower SNRs which, hamper the overall performance of the system to an extant.

There are certain other limitations such as impractical assumptions made about channel. In particular in [19] a deterministic blind method is presented, which requires the channel transfer functions to be coprime(no common zeros). There is an another restriction on the

input signal constellation also that, they have to have constant-modulus (CM). In other methods precoding is suggested for blind channel estimation, it should be noted that precoding not only increases system complexity but also consumes additional bandwidth.

The other problem with the conventional (O)STBC decoding is, it is extremely sensitive to the channel's estimation error. To demonstrate this effect, a STC system has been considered. This system has subspace based semi-blind channel estimation. Figure 1. shows the sensitivity of system performance to the channel estimation error.

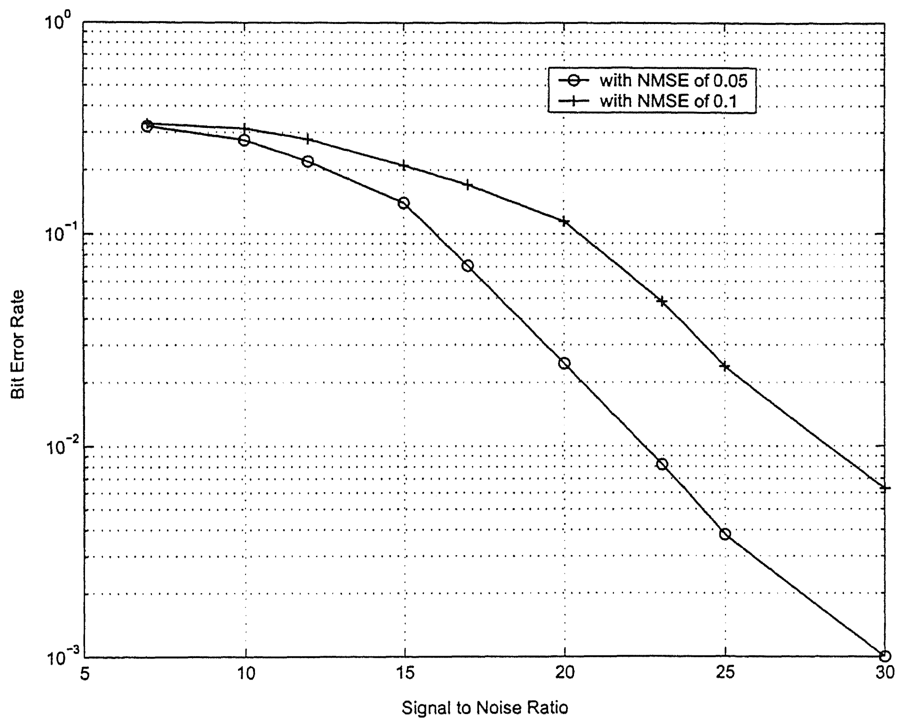

Figure 1.1: Plot showing the comparison of performance when estimation error Normal Mean Square Error(NMSE) is 0.05 and 0.1

The configuration of system considered for plots in Figure 1.1 is: 2 transmit and 1 receive antennas. Coding used is, Alamouti [6] OSTBC with OFDM modulation. Channel is non-flat faded but remains constant for the entire period of estimation and decoding. The channel estimation method is subspace based method [18] with appropriate number of pilot symbols to adjust NMSE of channel estimation.

It can be easily observed from the Figure 1.1 that the performance of system is very sensitive to the channel estimation error. The number of pilot symbols used For the two cases considered NMSE equal to 0.05 and 0.1 are almost 18 and 14 respectively. The complexity at the receiver for channel estimation is considerably high for both the cases. Even though the performance is comparable between at the lower SNRs. The distance between the curves is quite high at higher SNRs. For any further increase in the performance by reducing the NMSE with more pilot symbols, receiver not only becomes formidably complex but also more bandwidth is consumed.

1.5 Differential Orthogonal Space Time Block Coding

Conventional OSTBC techniques have the problems as explained, to be used for practical systems. To avoid this practical problems, already existing differential techniques for single antenna systems can be extended to MIMO case. Differential en/decoding techniques for MIMO case, which does not require any channel/frequency offset estimation has attracted considerable research attention in the recent past [12] [11].

This approach obviates or reduces training but incurs certain amount of performance loss. Nevertheless, the performance offered by this technique is exciting in adversely fading channel conditions.

For single transmit antenna systems, differential detection schemes exist that neither require the knowledge of the channel nor employ pilot symbol transmission. These differential decoding schemes are used, for instance, in the IEEE IS-54 standard. This motivated the generalization of differential detection schemes for the case of multiple transmit antennas. In [12] a truly differential detection technique was proposed, which requires practically no pilot symbols at all. Later in the same year B. M. Hochwald [24] proposed differential technique for unitary Space-Time codes. As these systems obviate the need to estimate the channel, the added advantage is that these promise consistent performance even when channel fades fastly, i.e., channel with much less coherence time.

Conventional Space-Time codes [6] and [8] assume that channel remain constant for large period of time. This time can be some times, as high as 20-22 transmission blocks time period. So, these systems results in poor performance when coherence time of channel is considerably low. In contrast, differential schemes [25] [26] require channel to be constant for considerably low period of time. The actual assumption over coherence time is usually a little above the 2 transmission blocks time period. All these attractive features of differential techniques come with a performance loss, which is up to 3 dB in some cases (the mathematical reason is shown for this in further chapters). Thus any technique to improve the performance in differential method is well appreciated.

Chapter 2

Amplitude PSK Modulation

2.1 Evolution of Amplitude PSK Modulation

APSK (Amplitude Phase Shift Keying) is a bandwidth efficient modulation technique. This was introduced way back in 1991 by Web [27]. Ever since, it is attracting the attention of researchers. Application of APSK technique to many standard encoding and decoding techniques resulted in higher throughput.

Mary APSK consists of two $M/2$ary PSK rings of . For example a 16 sized APSK constellation is shown Figure 2.1. This constellation does not have minimum least free distance between points in the strict sense. But this does allow efficient differential decoding methods to be used which go someway towards mitigating the effects of fading channel conditions.

2.1.1 Differential Detection with Amplitude PSK signals, for Single Antenna Systems

APSK when used with simple block codes (other than differential codes) facilitates transmission of an extra bit with each block of data. The block of data is modulated with PSK symbols and the amplitude of which contains the information about the extra bit of the block. The original PSK symbols amplitude switch between inner PSK ring radius and outer PSK ring radius or remain same, depending on the value of that last bit. The reverse of this is

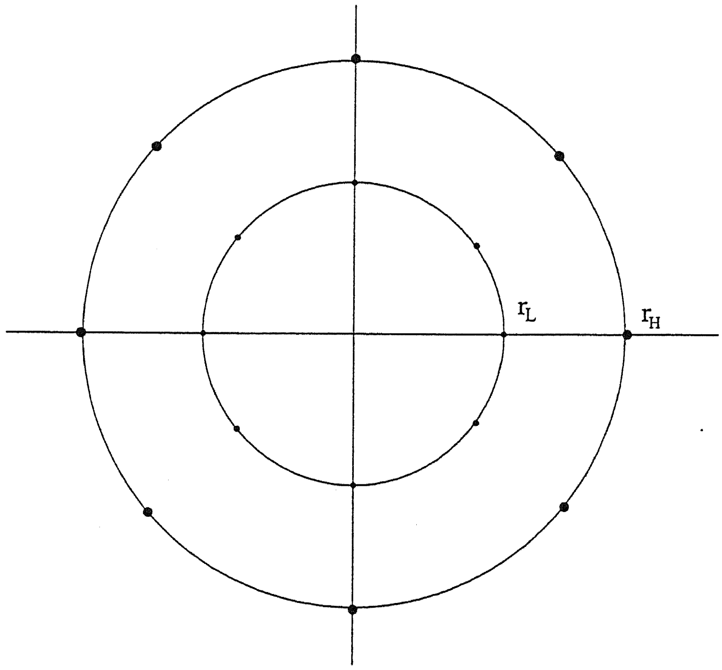

Figure 2.1: This is 16APSK constellation, which is essentially a pair of 8PSK rings with different radii(r_L and r_H)

done at the receiver. It just keeps track of the signal strength in general and decides last bit is '0' if there is no significant change in the amplitude and '1' if there is. In this manner, if APSK is used, always an extra bit can be sent with each block of data.

There is a problem when Amplitude PSK is used in the systems, where the fading coefficients are estimated for decoding. The amplitude modulated by this extra bit actually multiplies the fading coefficient and effectively it changes the fading coefficient. This causes the incorrect estimation of fading coefficient and thus the incorrect decoding/detection of the message symbols. This may also, lead to the error propagation to happen. This then, not only effect the decoding of this last extra bit but also effects the decoding of other PSK symbols in the block. This problem will never arise when we use the techniques, where we do not need to estimate channel at all. i.e., systems employing techniques like differential en/decoding. This is the reason why usually APSK is coupled with differential technique in practical systems.

Digital communication systems using 16-QAM(Quadrature Amplitude Modulation) and conventional receiver techniques have unacceptably high BER in Rayleigh fading environment. This is because square QAM constellations suffers from what is known as false lock positions at 26^o and 56^o. As a solution to this problem a star QAM (also known as APSK) has been proposed [27]. This was further refined in [28] to make it more robust for differential reception in the case of Rayleigh faded channels. Later this technique is refined to make it applicable for more practical channel conditions and with different coding techniques such as OFDM in [29].

System for 16 DAPSK: Constellation for this is as shown in the Figure 2.1 and we use the notation from the [31]. The constellation essentially consists of two independent 8 PSK rings $\exp(2\pi jm/8)$:$0 \leq m \leq 7$ and one 2 Differential Amplitude-Shift Keying (2DASK) r_L, r_H. where r_H and r_L are the inner and outer radii of the PSK constellations. It is shown in [31] that for optimal BER performance these r_H and r_L should satisfy following set of equations

$$r_L^2 + r_H^2 = \frac{1}{2} \tag{2.1}$$

$$r_H/r_L = a \tag{2.2}$$

$$r_l = \sqrt{\frac{2}{a^2+1}} \tag{2.3}$$

where 'a' can vary from 1.5 to 2. The four bit symbol (a_n, b_n, c_n, d_n) at the nth T seconds are differentially encoded and decoded, where the first three bits (a_n, b_n, c_n) are carried by the 8DPSK and the last bit d_n is carried by 2DASK. According to the value of this last bit, the proper PSK ring is chosen, for example if the bit d_n is 'zero' then we don't have change in the amplitude of the PSK ring, we use same ring as that was used in the previous frame. If d_n is 'one' then we switch between the two PSK rings.

Differential Detection rule: Here we shall discuss about the Maximum Likelihood detection rule for 16DAPSK symbols.

Let r_n be the present received symbol, given by:

$$r_n = A_n\sqrt{(\rho)}e^{j(\theta_n-\phi)} + n_n \tag{2.4}$$

where θ_n is the phase angle of the transmitted symbol at nth signalling duration, ϕ is the carrier phase and n_n is the complex AWGN and ρ is the signal strength. A_n is the amplitude, decided by the extra bit sent in each frame (noting that, this is Amplitude PSK), depending on the last bit of the frame this An can be one either r_L or r_H.

Similarly the previous received symbol r_{n-1} can be given as:

$$r_{n-1} = A_{n-1}\sqrt{(\rho)}e^{j(\theta_{n-1}-\phi)} + n_{n-1} \tag{2.5}$$

The detection of the bits involves two steps, first we decode the last bit of the frame. This is done by tracking the change of the amplitude change in the received signal. We calculate the strength (taking modulus of, mathematically) of the received signal. If there is a significant change in the amplitude of the signal, we can decide that last bit of the frame, d_n is 'one' and 'zero' if not.

Secondly we use phase detector to detect the PSK symbols. The decision variable for the phase detector is the phase difference between these two complex numbers. Equivalently, we can project r_n onto r_{n-1} and use the phase of the resulting complex number; that is,

$$r_n r_{n-1}^* = \rho e^{j(\theta_n-\theta_{n-1})} + \sqrt{(\rho)}e^{j(\theta_n-\phi)}n_{n-1}^* + \sqrt{(\rho)}e^{-j(\theta_{n-1}-\phi)}n_n + n_n n_{n-1}^* \tag{2.6}$$

which clearly in absence of noise, yields the phase difference $\theta_n - \theta_{n-1}$. This metric is then used to find the actual PSK symbol transmitted as follows:

The above calculated metric is multiplied with a PSK symbol and maximized over all possible PSK symbols of the constellation, to find actual transmitted symbol [50].

$$\hat{c} = \arg \max_{c \in \mathcal{S}_M} Re\{r_n r_{n-1}^*\} \tag{2.7}$$

where $\mathcal{S}_M$ is the M-ary PSK constellation used for modulation and $Re\{.\}$ is the 'real part of' operand. $\hat{c}$ is the detected PSK symbol.

2.1.2 Differential Detection of Amplitude PSK signals for MIMO systems

Once the potential capacities of MIMO channels were unveiled, the significance of spectrally efficient techniques such as APSK modulation drew attention of researchers. Application of APSK signalling to the high capacity channels like MIMO makes system even more spectrally efficient.

The key work in this area of APSK with OSTBC has been presented in [25]. Recently in [30] APSK has applied to unitary Space-Time codes. In particular, we shall deal with the technique proposed in [25], in further sections

2.2 Differential Detection of MPSK signals for Differential Orthogonal Space Time Block Codes

For a general spacetime code $\mathcal{C} = \{\mathbf{C}_i, i = 0, 1, 2, ..., M-1\}$ of M different $N \times N$ unitary matrices with $\mathbf{C}_i^\dagger \mathbf{C}_i = I$. The following differential spacetime coding was proposed in [32] for N transmit antennas:

$$\mathbf{S}_n = \mathbf{S}_{n-1}\mathbf{C}_{k_n}, \qquad n = 1, 2, ... \tag{2.8}$$

where $\mathbf{S}_0$ is an arbitrary fixed unitary matrix and $k_n$is encoded from the current $log_2(M)$ bits of an information sequence over the time duration of NT seconds. The transmitted signal $\sqrt{\rho}\mathbf{S}_n$ is with the signal power ρ. In the existing differential spacetime coding in [32] and [12], all the unitary matrices $\mathbf{C}$ have the same norm. If the orthogonal spacetime block codes, such as the Alamoutis code in [6], are used, the two symbols x_1 and x_2 used in a time block have

to satisfy that the sum of their powers, $|x_1|^2 + |x_2|^2$, has to be constant for any independent symbols x_1 and x_2 in a signal constellation. This implies that the signal constellation has to be a PSK, and may limit its performance when a high bandwidth efficient modulation scheme is preferred.

In our simulations, we are only interested in the comparison with the differential STC using the Alamoutis code in [6] and the PSK, which is named STC-PSK in short. Its detailed encoding is as follows.

Let$K = log_2(M)$. For $2K$ the bits of information $(I_1(n), ..., I_K(n), I_{K+1}(n)..., I_{2K}(n))$at the time $2nT$, the first K bits, $I_1(n), ..., I_K(n)$, are mapped to $x_1 \in \mathbf{S}_M$ and the second K bits, $I_{K+1}(n)..., I_{2K}(n))$, are mapped to $x_2 \in \mathbf{S}_M$. Then, a spacetime codeword matrix is formed as shown below:

$$C(x_1, x_2) = \frac{1}{\sqrt{(2)}} \begin{pmatrix} x_1 & -x_2* \\ x_2 & x_1* \end{pmatrix} \text{ where } x_1 \in S_{M_1} \text{and } x_2 \in S_{M_2} \tag{2.9}$$

Then, the transmitted signals for two antennas are the two rows of matrix $\sqrt{(\rho)}\mathbf{S}_n$ during the time intervals $[2(n-1)T, (2n-1)T]$ and $[(2n-1)T, 2nT]$, respectively, where $\mathbf{S}_0 = I$ and $\mathbf{S}_n = \mathbf{S}_{n-1}.\mathbf{C}(x_1, x_2)$, for $n = 1, 2, ...$ The simulation of this system in comparision with DOSTBC with APSK is shown later in this chapter.

2.3 Differential Detection of Amplitude PSK signals for Differential Orthogonal Space Time Block Codes

In this section, we analyse a differential en/decoding scheme for Alamoutis orthogonal space-time code using amplitude/ phase-shift keying (STC-APSK) signals. Antenna configuration is two transmit antennas and one receive antenna.

2.3.1 Encoding and Decoding methods

In this section we shall discuss the encoding and decoding methods of Differential OSTBC [6] with APSK (in particularly with 16APSK). The notation used is from [25]. We now describe STC-(M_1+M_2)DAPSK scheme. Its signal constellation consists of two independent M_1 PSK and M_2PSK and one 2ASK as shown in Figure 2.1 for $M_1 = 8$ and $M_2 = 8$. The relation between r_H and r_L and parameter a remains same as in the case for single antenna.

In each $2T$ seconds, two symbols x_1 and x_2 are transmitted and their amplitudes may be different in the next $2T$ seconds, and totally $(log_2(M_1) + log_2(M_2) + 1)$bits are carried in $2T$ seconds. Therefore, the bandwidth efficiency is $(log_2(M_1) + log_2(M_2) + 1)/2$b/s/Hz. Note that the bandwidth efficiencies for the MAPSK and the STC- M PSK are $log_2(M)$b/s/Hz only. The detailed encoding is as follows.

Differential Encoding:

Let $K_i = log_2(M_i)$ for $i = 1, 2$. For each $K_1 + K_2 + 1$ bits of information. i.e., the frame of bits $(I_1(n), ..., I_{K_1}(n), I_{K_1+1}(n), ..., I_{K_1+K_2}(n), I_{K_1+K_2+1}(n))$ during the time interval $[2(n-1)T, 2nT]$, the first K_1 bits, $I_1(n), ..., I_{K_1}(n)$, are mapped to $x_1 \in S_{M_1}$, the second K_2 bits, $I_{K_1+1}(n), ..., I_{K_1+K_2}(n)$, are mapped to $x_2 \in S_{M_2}$. The last bit $I_{K_1+K_2+1}(n)$ determines whether the mean signal power of S_n stands the same as the previous one S_{n-1} or change between r_L and r_H. The algorithm can be briefed mathematically as: $S_0 = a_0 P_0$, with $a_0 = r_L$ and $P_0 = I$

$a_n = a_{n-1} b_n$, where

$$b_n = \begin{cases} 1, & \text{if } I_{K_1+K_2+1}(n) = 0 \\ r_H/r_L, & \text{if } I_{K_1+K_2+1}(n) = 1 \text{ and } a_{n-1} = r_L \\ r_l/r_l, & \text{if } I_{K_1+K_2+1}(n) = 1 \text{ and } a_{n-1} = r_h \end{cases}$$

$P_n = P_{n-1}.C(x_1, x_2)$

$S_n = a_n P_n$. for $n = 1, 2, ...,$

where $C(x_1, x_2)$ is the 2×2 space-time codeword matrix formed by Alamouthi STC, given by: 2.9.

This encoding technique is also illustrated in a flowchart in Figure 2.2

Differential Decoding:

Consider single receive antenna at the receiver. The received signals in the time intervals

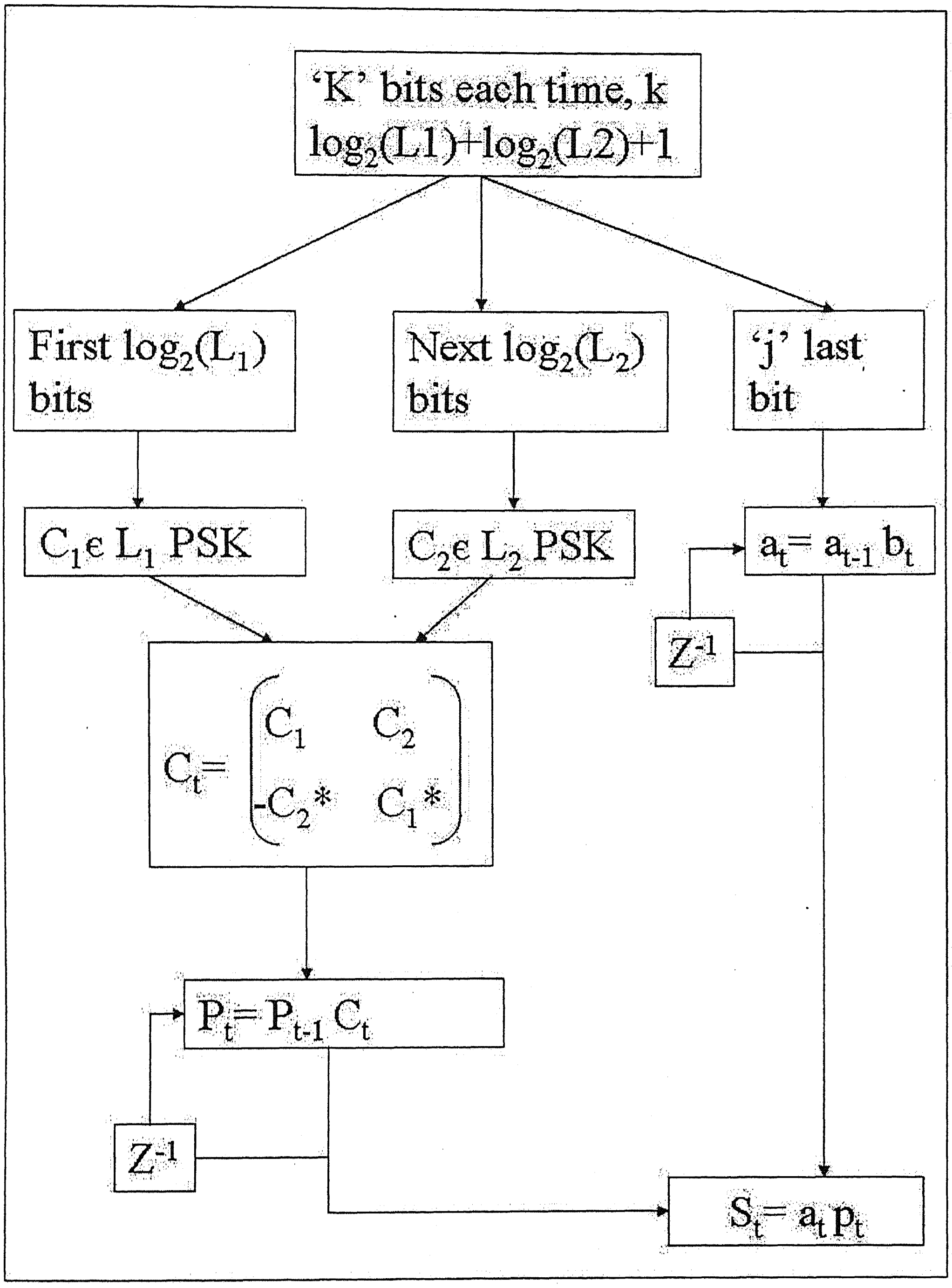

Figure 2.2: Flowchart explaining the Differential Orthogonal Space-Time Block Codes encoding.

$[2(n-1)T, (2n-1)T]$ and $[2(n-1)T, 2nT]$ form the following 1×2 vector:

$$\mathbf{r}_n = \sqrt{\rho}\mathbf{h}\mathbf{S}_n + \mathbf{w}_n \tag{2.10}$$

where $\mathbf{h} = [h_1, h_2]$, h_1 and h_2 are the channel gains or coefficients from the first and the second transmit antennas to the receive antenna, respectively, and they are independent complex Gaussian random variables of mean 0 and variance 1, and $\mathbf{w_n} = [w_{n,1}, w_{n,2}]$ is the additive complex Gaussian noise vector with two independent identically distributed components of mean 0 and variance N_0. In our simulations, this variance is determined by the total energy per bit, E_b at the transmitter over N_0 i.e., E_b/N_0.
From the method of encoding and (2.10), the received signal vector can be rewritten as:

$$\mathbf{r}_n = \mathbf{r}_{n-1} b_n \mathbf{C}(x_1, x_2) + \hat{\mathbf{w}}_n \tag{2.11}$$

where

$$\hat{\mathbf{w}}_n = \mathbf{w}_n - \mathbf{w}_{n-1} b_n \mathbf{C}(x_1, x_2). \tag{2.12}$$

From the equation (2.11) it is evident that our decoding scheme has the noise as $\hat{\mathbf{w}}_n$ instead of simple $\mathbf{w}_n$. The variance and thus the noise power of $\hat{\mathbf{w}}_n$ from (2.12) is double to that of simple $\mathbf{w}_n$. This explains the performance loss of differential technique when compared to the conventional OSTBC decoding in general. A plot showing the comparison of performance between the differential, conventional channel estimation technique and the (ideal) coherent reception has been provided in the further chapters.

The differential decoding then has two steps. The first step is to detect the (K_1+K_2+1)th bit by detecting whether b_n is 1 or not: by noticing that $b_n \in 1, a, 1/a$. This can be done by simply tracking the amplitude of the signal, this is very similar to the single antenna case except in the place of modulus of the received symbol, we take the Frobenius norm of the received matrix. This can be mathematically expressed as:

$$\hat{b}_n = \arg \min_{b \in \{1,a,1/a\}} |||\mathbf{r}_n|| - b||\mathbf{r}_{n-1}||| \tag{2.13}$$

where $||\mathbf{V}|| = \sqrt{\sum_i |v_i|^2}$, the Frobenius norm if $\mathbf{v}= (v_i)$. If $\hat{b}_n = 1$, then $\hat{I}_{K_1+K_2+1}(n) = 0$, otherwise $\hat{I}_{K_1+K_2+1}(n) = 1$.
After b_n is detected, the second step is to detect the PSK symbols x_1 and x_2 for the first K_1

bits and the second K_2 bits, respectively. The method shown here is the result of simplification of Maximum Likelihood (ML) detection technique.

$$(\overline{x}_1, \overline{x}_2) = \arg \min_{x_1 \in S_{M_1}, x_2 \in S_{M_2}} ||\mathbf{r}_n - \overline{b}_n \mathbf{r}_{n-1} \mathbf{C}(x_1, x_2)||^2 \tag{2.14}$$

Let the two components of received vector r_n be defined as: $r_n = [r_{n,1}, r_{n,2}]$;
The actual mathematical equation used for decoding can be derived simply from the case of single antenna, which was clearly explained in the first section of this chapter. By considering the Orthogonality of the two channels the first PSK symbol x_1 can be detected as follows:

The symbol x_1 is sent twice over the orthogonal channels, so we form two metrics. Each of which is in line with the single antenna case. The first metric would be: $\mathbf{r}^*_{n,1}\mathbf{r}_{n-1,1}$ considering the symbol sent in the first time slot. In the next time slot the symbol is conjugated before sending, so the metric from second time slot will be $\mathbf{r}_{n,2}\mathbf{r}^*_{n-1,2}$. Since the channels of the 2×1 MIMO STBC system are proved to be orthogonal(in the section 1.3.1), we can add these matrices for the detection of the symbol x_1. There fore the detection rule for the symbol x_1 is:

$$\hat{x}_1 = \arg \max_{x_1 \in S_{M_1}} \mathbf{Re}\{(\mathbf{r}^*_{n,1}\mathbf{r}_{n-1,1} + \mathbf{r}_{n,2}\mathbf{r}^*_{n-1,2})x_1\} \tag{2.15}$$

Where **Re** stands for the 'real part of' operand.
Next for the second PSK symbol x_2, we can use the same detection rule as that of x_1 with slight change, considering the position of the symbol x_2 in the initial data matrix $\mathbf{S}_n$. Also, this x_2 is negated in addition to the conjugation for the second time slot. So, the metrics for detection of x_2 would be: $\mathbf{r}^*_{n,1}\mathbf{r}_{n-1,2}$ and $-\mathbf{r}_{n,2}\mathbf{r}^*_{n-1,1}$. Now, as the two channels of the system are orthogonal, so we can add these two metrics for final detection of x_2.

Therefore, the detection rule for x_2 is:

$$\hat{x}_2 = \arg \max_{x_2 \in S_{M_2}} \mathbf{Re}\{(\mathbf{r}^*_{n,1}\mathbf{r}_{n-1,2} - \mathbf{r}_{n,2}\mathbf{r}^*_{n-1,1})x_2\} \tag{2.16}$$

It can be easily observed that the complexity of the system is very much similar to that of single antenna case[28].

2.3.2 Simulation Results and Conclusions

In simulations, E_b/N_0 is used as the channel SNR, where E_b is the total energy per bit used in the transmission and is the summation of energies in all transmit antennas in a multiple antenna system. In all simulations, 100 information symbols are used in each trial and 10,000 trials are used. The fading channels are flat, i.e., constant, in each trial and are complex Gaussian random variables over 10,000 trials. Two different kinds of differential en/decoding schemes are compared. The first one is the 16-DPSK with multiple transmit antennas and one receive antenna (DPSK-STBC). The bandwidth efficiency here is 4 b/Hz/sec. The second system is again a MIMO system with two transmit and one receive antennas, modulation is APSK (DOSTBC-APSK system). The bandwidth efficiency here is 4.5 b/Hz/sec. The channel considered for the simulations is flat faded Rayleigh channel and modelled by modified jake's model (discussed in detail in chapter 4). The channel coherence time, is taken to be equal to the time taken for two blocks of data transmission.

It can be seen from the plot of Figure 2.3 that the performance of the two systems are comparable with DOSTBC-PSK being slightly better. The conclusion is DOSTBC-APSK system has a higher bandwidth efficiency at the cost of very little degradation in the performance.

Next we consider the effect of size of constellation on the OSTBC-DD system's performance. As the constellation size increases the bandwidth efficiency of the system goes up. For example, 32APSK it is 5.5 bits/sec/Hz, for 16APSK it is 4.5 bits/sec/Hz and for 8APSK it is 3.5 bits/sec/Hz. This increase in spectral efficiency is at the cost of the performance loss of the overall system. When the size of the constellation increases, the average distance between the symbols decrease. This causes the average error probability to come down. So the spectral efficiency and the system performance are to be traded-off properly. In the Figure 2.4 the DOSTBC-DD system is evaluated with 32APSK, 16APSK and 8APSK. We can see that spectral efficiency of the 32APSK system is higher than others but the performance is worse than others.

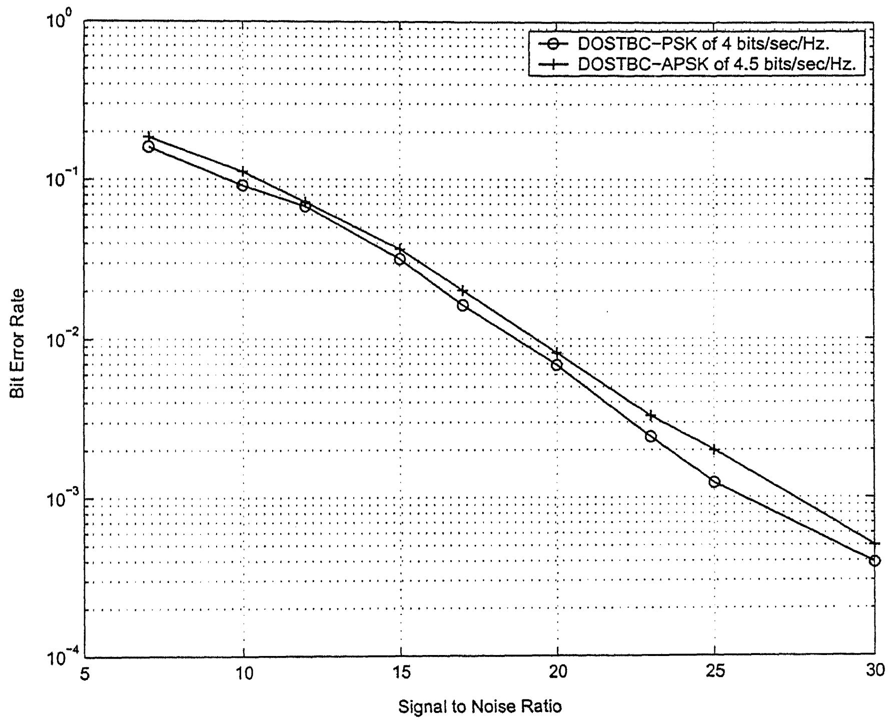

Figure 2.3: Figure showing the comparison of performance between DOSTBC-APSK and DOSTBC-PSK systems

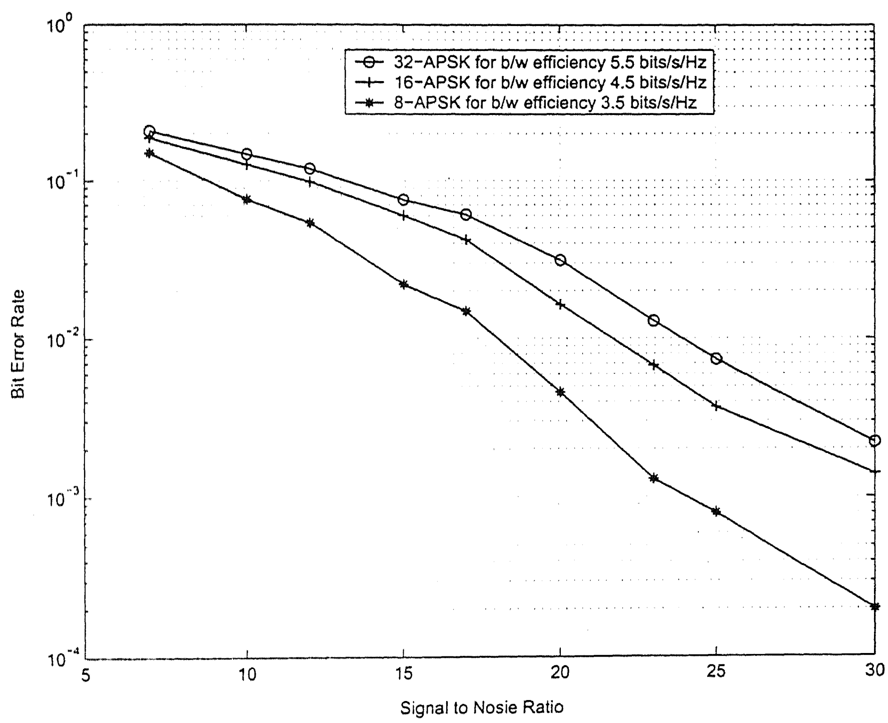

Figure 2.4: Figure showing the effect of change of constellation on the performance of of the DOSTBC APSK system

Chapter 3

Decision Feedback Scheme with Amplitude PSK

As discussed in the earlier chapters, the Differential OSTBC system with APSK is bandwidth efficient scheme. It has been shown that the DOSTM-APSK of 4.5 b/s/Hz performs approximately the same as the DOSTM with PSK signals of 4 b/s/Hz in terms of the bit-error rate (BER) versus E_b/N_0. In other hand the performance of the system needs to be improved when compared to their non-differential counterparts.

Over rapid-fading channels, differential detection for differential schemes (either for single transmit antenna or for multiple transmit antennas) has an error floor [33]. In the single transmit antenna case, this problem is well studied and a lot of work has been done for differential PSK modulation [34], [35] and for the differential APSK modulation [36] [37] [38]. The error floor problem becomes more serious in the multiple transmit antenna systems since the differential schemes assume the channel coefficients constant in adjacent blocks. When the block length or the number of tranmist antennas increases, the assumption is more vulnerable. Decision Feedback Differential Detection(DF-DD) technique for multiple block detection was proposed in [39]. But that is based on the second order statics of the channel, which are not always available. In [40] a DF-DD scheme for OSTBC is also proposed (which we discuss in detail in the later parts of this chapter), where channel equalizer principle is made use for differential detection.

3.1 Optimal Coefficient method for Decision Feedback Equalization for Flat fading channels

The encoding and decoding algorithms of DOSTM technique had been already discussed in detail in previous chapter. Here we will have the encoder at transmitter as unchanged, making the receiver more optimal. In the earlier decoding technique we assumed that channel $\mathbf{h}$ remains constant for adjacent blocks of data transmission. The decoder uses the previous received symbol $\mathbf{r}_{n-1}$ as an estimation of $\sqrt{\rho}\mathbf{h}\mathbf{S}_{n-1}$ as shown in [33] for the detection of current information symbols $\mathbf{b}_n$ and $\mathbf{C}_n$. Where as in reality the channel vector $\mathbf{h}$ is not constant but of fading. In this case, we can naturally use more received data instead of just last received symbol, to compensate for fading of the channel. Here in this section we shall discuss the linear prediction based DF-DD for the DOSTM-APSK.

In the DF-DD, the past $N-1$ block received signals $\mathbf{r}_{n-1}, \mathbf{r}_{n-2}, ..., \mathbf{r}_{n-N+1}$ are used to estimate $\sqrt{\rho}\mathbf{h}\mathbf{S}_{n-1}$. In particular, the following $\mathbf{r}_{n-1}^{ref}$, a linear combination of $\mathbf{r}_{n-1}, \mathbf{r}_{n-2}, ..., \mathbf{r}_{n-N+1}$ is used as an estimate of $\sqrt{\rho}\mathbf{h}\mathbf{S}_{n-1}$:

$$\mathbf{r}_{n-1}^{ref} \triangleq \sum_{v=1}^{N-1} g_v \mathbf{r}_{n-v} \tag{3.1}$$

where g_v $v = 1, 2, ..., N-1$, are parameter coefficient matrices to be determined and N is defined as the feedback order of the DF-DD. Since noise assumed, is Gaussian noise and independent of $\mathbf{s}_{n-1}\mathbf{h}_n$, the parameter coefficient matrices g_v can be solved by minimizing the following mean square error (MSE):

$$J = E\Big\{\|\mathbf{r}_n - \mathbf{b}_n\mathbf{c}_n \sum_{v=1}^{N-1} g_v \mathbf{r}_{n-v}\|^2\Big\} \tag{3.2}$$

Next the main goal of the analysis would be finding out a more optimal solution for parameter coefficient matrices g_v. But before going any further into that, we have to make some observations on these parameter coefficient matrices.

1)Channel has certain correlation across the time domain, because it is Rayleigh faded channel.

2) Information data sequence is I.I.D.(Independent and Identically Distributed), so they do not have contribution towards the correlation between the received symbols.

Thus the to be found parameter coefficient matrices g_v should have the forms such that the information data sequence can be compensated (roughly) in the objective function in the (3.2). Therefore, by properly choosing the parameter coefficient matrices, the objective function in (3.2) is independent (roughly) of the information data sequence $\mathbf{b}_n\mathbf{C}_n$, but depends on the channel vector $\mathbf{h}_n$ and the additive noise vector $\mathbf{w}_n$. Since the vector components of $\mathbf{h}_n$ are independent and the vector components of $\mathbf{w}_n$ are also independent, and these two vectors are independent each other, the parameter coefficient matrices g_v after the information data compensation should be scalers rather than matrices, which will be seen in details in the following.

Based on the above observations and the encoding scheme explained in chapter 2, we let

$$g_v = q_v \prod_{\mu=1}^{v-1} \mathbf{b}_{n-\mu}\mathbf{c}_{n-\mu} \tag{3.3}$$

where $q_v, v = 1, 2, ..., N-1$, are scalars and are determined later. Then from (3.2) and (3.3) we have

$$\begin{aligned} J &= E\Big\{\Big|\Big|\mathbf{r}_n - \sum_{v=1}^{N-1} q_v(\prod_{\mu=0}^{v-1} \mathbf{b}_{n-\mu}\mathbf{C}_{n-\mu})\mathbf{r}_{n-v}\Big|\Big|^2\Big\} & (3.4) \\ &= \rho||\mathbf{s}_n||^2.E\Big\{\Big|\Big|\mathbf{h}_n + \frac{\mathbf{p}_n^H}{a_n\sqrt{(\rho)}} - \sum_{v=1}^{N-1} q_v(\mathbf{h}_{n-v} + \frac{\mathbf{p}_{n-v}^H}{a_{n-v}\sqrt{\rho}}\mathbf{w}_{n-v})\Big|\Big|^2\Big\} & (3.5) \end{aligned}$$

In (3.5), all the information data terms have been absorbed into $\mathbf{r}_n$ such that $\mathbf{s}_n$ becomes the common term when g_v has the form in (3.3) and therefore, the common term $\mathbf{s}_n$ of the all the information data can be separated from the test as shown in (3.5). Therefore, the scalar parameter coefficients $q_v = 1, 2, ..., N-1$ are the coefficients of the forward linear prediction for the process.

$$\mathbf{f} = [f_n^1\ f_n^2]^T \triangleq \mathbf{h}_n + \frac{\mathbf{p}_n^H}{a_n\sqrt{\rho}}\mathbf{w}_n. \tag{3.6}$$

In (3.6), $\mathbf{p}_n$ is a unitary matrix, and two noise components in $\mathbf{w}_n$ are independent Gaussian noises, and two channel components in $\mathbf{h}_n$ are also independent and both of them have the same distribution. Therefore, the two components f_n^1 and f_n^2 are independent and identically distributed. This explains why coefficients, $q_v, v = 1, 2, ..., N-1$, of the forward linear prediction for the process are set to be scalars. From the orthogonality of matrices $\mathbf{p}_n$, the

autocorrelation of f_n^1 or f_n^2 is

$$r_f(\tau) = r_h(2\tau) + \frac{1}{2\rho}\left(\frac{1}{r_L^2} + \frac{1}{r_H^2}\right)\delta(\tau) \tag{3.7}$$

where $r_h(\tau) = E\{h_{m,n}h^*_{m,n-\tau}\}$ is the autocorrelation function of the channel $h_{m,t}$ and $\delta(\tau)$ is the Kronecker function. Thus, for the forward linear prediction problem [41], the minimum MSE (MMSE) solution can be given as:

$$q^* = \phi^{-1}\varphi \tag{3.8}$$

where

$$\begin{aligned}
\mathbf{q} &= [q_1\ q_2\ \ldots\ q_{N-1}]^T \\
\phi &= \begin{bmatrix} r_f(0) & r_f(1) & \ldots & r_f(N-2) \\ r_f(-1) & r_f(0) & \ldots & r_f(N-1) \\ \ldots & \ldots & \ldots & \ldots \\ r_f(-N+2) & r_f(-N+1) & \ldots & r_f(0) \end{bmatrix}. \\
\varphi &= [r_f(-1)\ r_f(-2)\ \ldots\ \ldots r_f(0)]^T.
\end{aligned}$$

It can be seen from all the above analysis that, the solution of parameter vector $\mathbf{q}$ in (3.8) needs the autocorrelation function of the channel. This technique of equalization is optimal when channel statistics are available at the receiver. Thus the name Optimal Coefficient method. But, in practice channel statistics are not always available at the receiver. We have to use the Adaptive algorithm such as RLS, LMS etc.

3.2 Recursive Least Square based Adaptive filter Equalization for Decision Feedback-Differential Detection scheme

As we have seen in the above section we cannot have the optimal solution for $\mathbf{q}$, the filter coefficients when the channel statistics are not available to us. To overcome this situation we go for adaptive filters at the receiver, which may or may not require initial training. In this

section we analyze the Root Least Squares(RLS) Algorithm based adaptive filter technique [40].

The way we minimize the MSE in 3.2 changes now with $\bar{\mathbf{r}} - i$ in place of $\mathbf{r}$, where:

$$\bar{\mathbf{r}}_{n-v} = \left(\prod_{\mu=0}^{v-1} \mathbf{b}_{n-\mu}\mathbf{c}_{n-\mu} \right) \mathbf{r}_{n-v} \tag{3.9}$$

Then according to [41] the cost function:

$$J_{LS}(\tau) = \sum_{k=1}^{\tau} \lambda^{\tau-k} \left|\left| \mathbf{x}_k - \sum_{v=1}^{N-1} q_v \bar{\mathbf{r}}_{k-v} \right|\right|^2 \tag{3.10}$$

where $0 < \lambda \leq 1$ is the forgetting factor. Following is the RLS algorithm that is applied here, to obtain $q_v, v = 1, 2, ..., N-1$, [41]

$$\begin{aligned}
\mathbf{q}_\tau &= \mathbf{q}_{\tau-1} + \mathbf{k}_\tau^* \xi_\tau & (3.11)\\
\mathbf{k}_\tau &= \frac{\lambda^{-1}\mathbf{v}_{\tau-1}\mathbf{u}_\tau}{1 + \lambda^{-1}\mathbf{u}_\tau^H \mathbf{v}_{\tau-1}\mathbf{u}_\tau} & (3.12)\\
\xi_\tau &= \mathbf{r}_\tau(1) - \mathbf{q}_{\tau-1}^T \mathbf{u}_\tau & (3.13)\\
\mathbf{v}_\tau &= \lambda^{-1}\mathbf{v}_{\tau-1} - \lambda^{-1}\mathbf{k}_\tau \mathbf{u}_\tau^H \mathbf{v}_{\tau-1} & (3.14)\\
\mathbf{u}_\tau &= [\bar{\mathbf{r}}_{\tau-1}(1)\ \bar{\mathbf{r}}_{\tau-2}(1)\ ...\ \bar{\mathbf{r}}_{\tau-N+1}(1)]^T & (3.15)
\end{aligned}$$

where $\mathbf{v}_0 = \sigma^{-1}\mathbf{I}_{N-1}$ and $\sigma = 0.004$ as suggested in [40]

The initial value of $\mathbf{q}_\tau$ can be set as $\mathbf{q}_0 = [1\ 0\ ...\ 0]^T$, which makes the detector be the original differential detector in [25]. To obtain the optimal value of $\mathbf{q}$, the data sequence $b_{n-\mu}\mathbf{c}_{n-\mu}$ in (3.9) can be assumed known or be replaced by the previous detected values $\hat{b}_{n-\mu}\hat{\mathbf{c}}_{n-\mu}$. If, in the beginning $b_{n-\mu}\mathbf{c}_{n-\mu}$ are assumed known, a training sequence is required. otherwise the system starts to work blindly.

Once the parameter coefficient matrices $\mathbf{q}$ are obtained, the estimate of $\sqrt{(\rho)}\mathbf{s}_{n-1}\mathbf{h}_n$ is

$$\mathbf{r}_{n-1}^{ref} = \sum_{v=1}^{N-1} q_v \left(\prod_{\mu=1}^{v-1} \hat{b}_{n-\mu}\hat{\mathbf{c}}_{n-\mu} \right) \mathbf{r}_{n-v}. \tag{3.16}$$

Then, the initial detection rule for the DOSTM-APSK can still be applied in the following steps.

step 1) Detect b_n from the metric

$$\hat{b}_n = \arg \min_{b \in \{1,a,1/a\}} \Big| \parallel \mathbf{r}_n \parallel - b \|\mathbf{r}_{n-1}^{ref}\| \Big|. \tag{3.17}$$

step 2) Detect c_1 and c_2 of the nth block from the metric

$$\hat{c}_i = arg \max_{c_i \in S_{M_i}} f_i(x_i), \qquad i = 1, 2. \tag{3.18}$$

where

$$f_1(c_1) = \mathbf{Re}\Big\{(\mathbf{r}_n(1)^* \mathbf{r}_{n-1}^{ref}(1) + \mathbf{r}_n(2)(\mathbf{r}_{n-1}^{ref}(2))^*)c_1\Big\} \tag{3.19}$$

$$f_2(c_2) = \mathbf{Re}\Big\{(\mathbf{r}_n(1)^* \mathbf{r}_{n-1}^{ref}(2) - \mathbf{r}_n(2)(\mathbf{r}_{n-1}^{ref}(1))^*)c_2\Big\} \tag{3.20}$$

It can be easily seen that these decoding equations differ from that of previous chapter. The $\mathbf{r}_{n-1}$ is being replaced by $\mathbf{r}_{n-1}^{ref}$, which is estimated from the N_1 previous received symbols and previously detected information symbols, rather than on only one previous symbol.

3.3 Modified Decision Feedback-Differential Detection technique

In this work, we suggest to modify the DF-Differential Detection technique slightly to get better performance. This modification is found to be giving better result compared to the original one, from literature [40]. The modification is made in the detection of the last bit of the frame, by more appropriate detection of b_n. The details of the modification made, is discussed with mathematical details below. It is also, supported by the simulation results shown in later part of this chapter.

3.3.1 Modification of Decision Feedback-Differential Detection detection technique for last bit of the frame

Consider the DF-DD detection equation for b_n (3.17), which is reproduced here for convenience:

$$\hat{b}_n = \arg \min_{b \in \{1,a,1/a\}} \Big| \parallel \mathbf{r}_n \parallel - b \|\mathbf{r}_{n-1}^{ref}\| \Big|. \tag{3.21}$$

Here we are minimizing the metric to find the proper value for $\hat{b}_n$, where $\mathbf{r}_{n-1}^{ref}$ can be written as:

$$\mathbf{r}_{n-1}^{ref} = \mathbf{r}_{n-1} + q_2\hat{b}_{n-1}\hat{c}_{n-1}\mathbf{r}_{n-2} + ... + q_{N-1}\left(\prod_{\mu=1}^{N-2}\hat{b}_{n-\mu}\hat{c}_{n-\mu}\mathbf{r}_{n-N+1}\right) \tag{3.22}$$

Consider the metric which was given for simple Differential Detection in chapter 2. Reproduced here for the convenience:

$$\hat{b}_n = \arg \min_{b\in\{1,a,1/a\}} \Big| \parallel \mathbf{r}_n \parallel - b\|\mathbf{r}_{n-1}\|\Big|. \tag{3.23}$$

Consider the case where noise is zero then, $\mathbf{r}_n = b_n c_n \mathbf{r}_{n-1}$ and the c_n is:

$$c_n = \frac{1}{\sqrt(2)}\begin{pmatrix} C_1 & C_2 \\ -C_2^* & C_1^* \end{pmatrix} \tag{3.24}$$

where c_i $i = 1, 2$ are PSK symbol with magnitude as unity, so the matrix c_n is such that:

$$||c_n A \parallel = \parallel A \parallel \tag{3.25}$$

where $\parallel . \parallel$ is Frobenius norm.

From this it is evident that metric in the equation (3.23) is becoming zero, leading to perfect recovery of b_n. Now consider the detection equation of b_n DF-DD scheme (3.17). By definition of $\mathbf{r}_{n-1}^{ref}$ in (3.22), the metric in (3.17) is non-zero, as the training coefficients $q_2, ...q_{N-1}$ are not all zero. So in noise less case the metric in (3.23) is minimal (in fact zero) compared to the metric in (3.17). This leads to the conclusion that, even in non-zero noise case the same remains minimum. So the optimal metric to be chosen for detection of b_n should be that of (3.23) rather than (3.17).

Now the consolidated modified decoding technique can briefed as:

step 1) Detect b_n from the metric

$$\hat{b}_n = \arg \min_{b\in\{1,a,1/a\}} \Big| \parallel \mathbf{r}_n \parallel - b\|\mathbf{r}_{n-1}\|\Big|. \tag{3.26}$$

step 2) Detect c_1 and c_2 on the nth block from the metric:

$$\hat{c}_i = arg \max_{c_i \in S_{M_i}} f_i(x_i), \qquad i = 1, 2. \tag{3.27}$$

where

$$f_1(c_1) = \mathbf{Re}\Big\{(\mathbf{r}_n(1)^*\mathbf{r}_{n-1}^{ref}(1) + \mathbf{r}_n(2)(\mathbf{r}_{n-1}^{ref}(2))^*)c_1\Big\} \tag{3.28}$$

$$f_2(c_2) = \mathbf{Re}\Big\{(\mathbf{r}_n(1)^*\mathbf{r}_{n-1}^{ref}(2) - \mathbf{r}_n(2)(\mathbf{r}_{n-1}^{ref}(1))^*)c_2\Big\} \tag{3.29}$$

3.4 Least Mean Square based Adaptive Equalization

In the DF-DD scheme the adaptive algorithm used is Recursive Least Squares (RLS), Which is efficient but has considerable implementation complexity. In this work, we have suggested to implement the same scheme by using much simpler Least Mean Square (LMS) in place of RLS. This modification not only reduced the system complexity considerably but also improved the system performance under certain situations. The following is the details of LMS adaptive algorithm in our case [41]:

$$\mathbf{q}_\tau = \mathbf{q}_{\tau-1} + \mu\xi^*\mathbf{u}_\tau \tag{3.30}$$

$$\xi_\tau = \mathbf{r}_\tau(1) - \mathbf{q}_{\tau-1}^T\mathbf{u}_\tau \tag{3.31}$$

$$\mathbf{u}_\tau = [\bar{\mathbf{r}}_{\tau-1}(1)\ \bar{\mathbf{r}}_{\tau-2}(1)\ \ldots\ \bar{\mathbf{r}}_{\tau-N+1}(1)]^T \tag{3.32}$$

The only new parameter in these set of equations is μ and it is the step size of the LMS algorithm. The value of this step size, μ is set by trail and error basis.

With this modification in the adaptive algorithm in DF-DD scheme, it has been observed LMS-DD is better performing than DF-DD in following cases:

Case i) When the channel is slowly fading, that is coherence time of channel is more than one data transmission block duration.

Case ii) When the order of the adaptive algorithm, N is less than 5. The proofs for all the claims made, are shown later in the simulation results section.

3.5 Simulation results and Conclusions

The simulation results and some conclusions are presented in this section for the DF-DD of DOSTM-APSK schemes. The system under consideration is 2 transmit antennas and 1

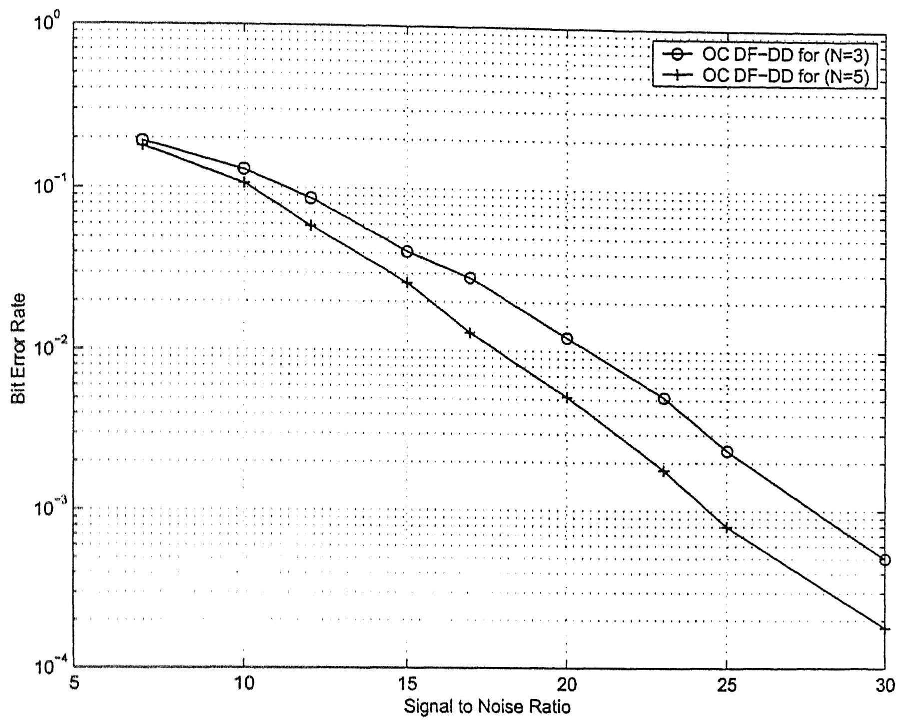

Figure 3.1: OC DF-DD performance for DOSTM-16APSK in the slow fading channel model, $f_d T_s = 0.015$

receive antenna with DOSTM -16APSK signalling. Gray mapping is employed for 8PSK symbols c_1 and c_2. The RLS equalization is referred to as adaptive DF-DD. In the adaption process, 200 iterations via a training sequence are used.

The channel has been modelled by using modified jake's model for Rayleigh fading channel [42] and [43]. Simulations were run on 2 different channels. One is slow fading (or block fading) channel, where channel coherence time is equal to at least 2 blocks of data transmission. The autocorrelation function of this channel can be give as: $r_h(t) = J_0(4\pi f_d T_s t)$. The other is fast fading (continuous fading) channel, where the coherence time is equal to one block of data transmission. The autocorrelation function is given by $r_h(t) = J_0(2\pi f_d T_s t)$. In both the channels, the fading rate is set to be $f_d T_s = 0.015$. The DF-DD enhances the detection performance of DOSTM-16APSK in both channel models. The performance is better when the feedback order N in the DF-DD increase.

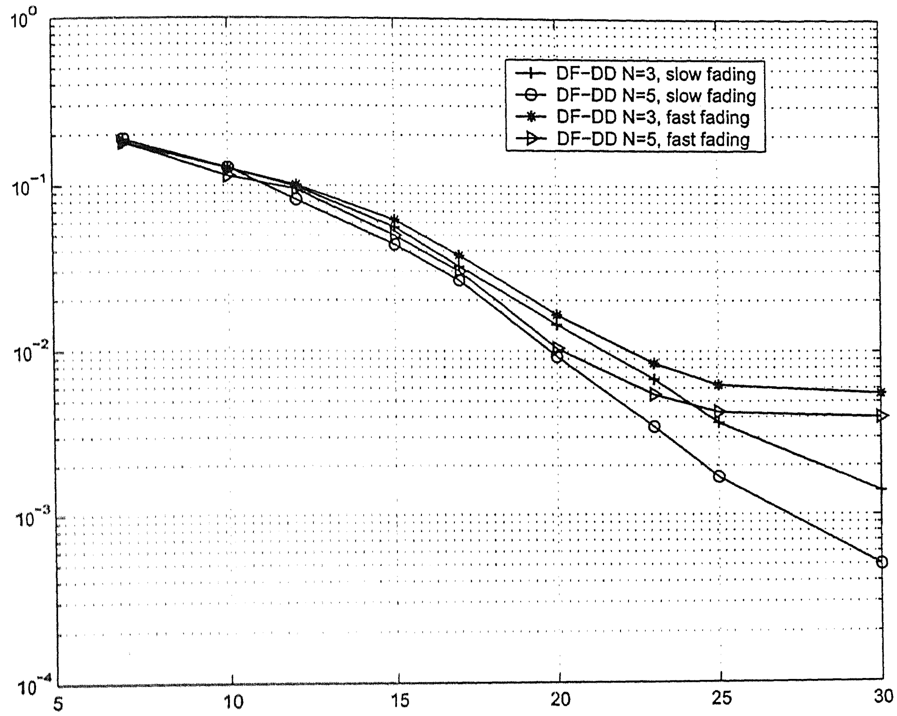

Figure 3.2: DF-DD performance for DOSTM-16APSK in the slow and fast fading channel models for 2 different values of N (the order of the adaptive algorithm),

First in Figure 3.1, we show the comparison of performance when the system has order of adaptive filter N as 3 and 5. The channel taken is slow fading (which is defined above).

Next in Figure 3.2, we compare the performance of the system in two different types of channels, i.e., slow and fast fading channels and with two different orders of the adaptive algorithm used in DF DD. It can be observed that the performance of the system is better in slow fading channel case compared to that of fast fading case for any particular order(N). Also, the performance of system is better if the order of the adaptive algorithm is increasing from $N = 3$ to $N = 5$, for any particular (either slow or fast fading) channel.

Next we consider the effect of changing the the initialization values of RLS algorithm based adaption scheme. The parameter σ of the algorithm explained in section 3.2 is the initialization of autocorrelation matrix $\mathbf{v}$, the performance of the algorithm is sensitive to this parameter. In the figure 3.3 we show the performance of this RLS based adaptive scheme

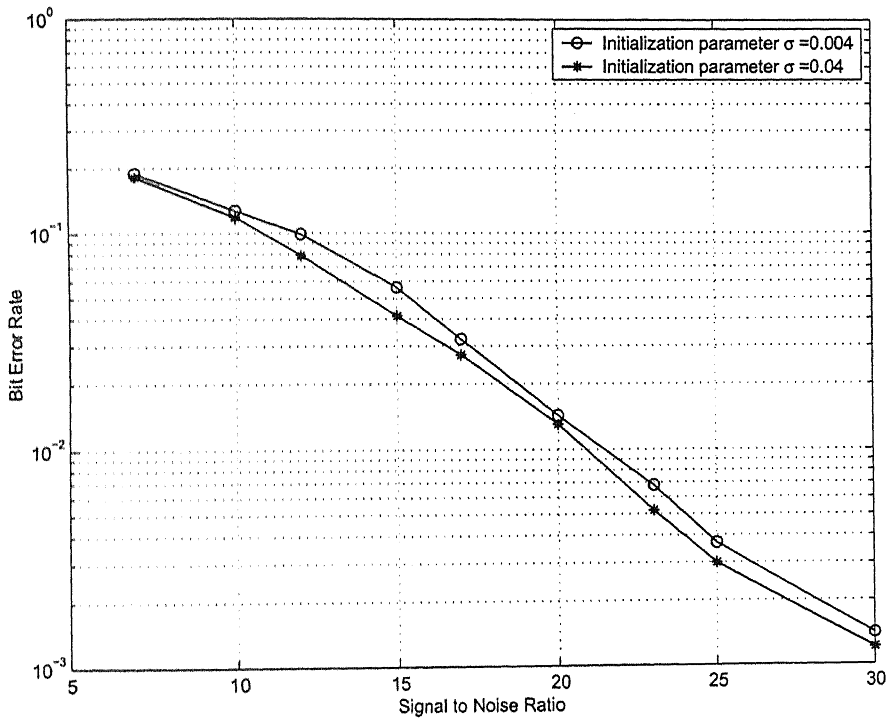

Figure 3.3: Effect of change of initialization parameter σ in RLS based DF-DD scheme. Channel is slowfading and the order of algorithm N is 3.

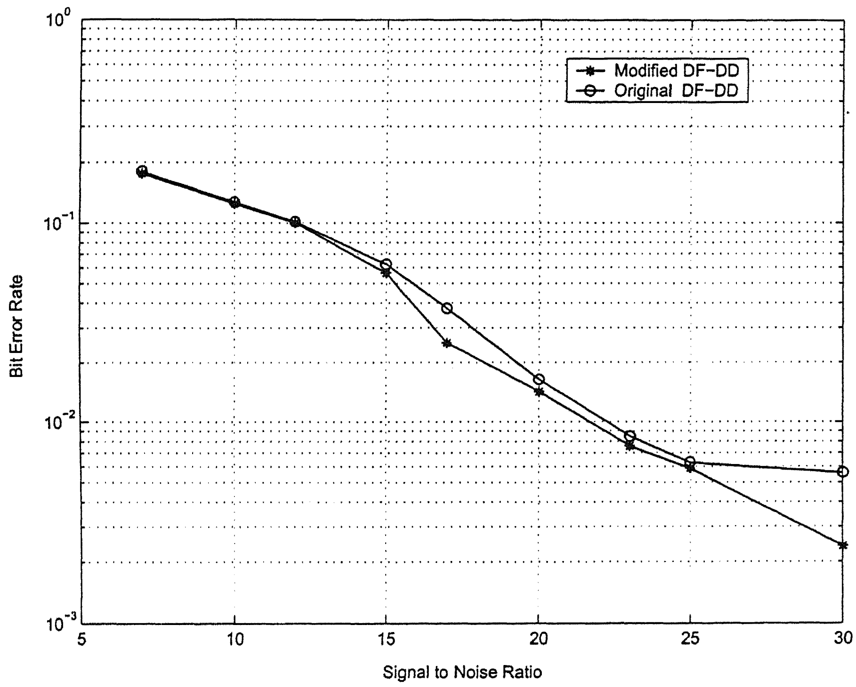

Figure 3.4: Comparison of system performance when modified DF-DD is applied and when original DF-DD [40] is applied in fast fading channel

with two different values (0.004 and 0.04) for σ (initialization of $\mathbf{v}$). The channel considered is slowfading channel and the order of the algorithm N is 3.

Next in the the Figure 3.4, we show the comparison of performance between the systems when our modified DF-DD is applied and when [40] is applied. The channel considered is fast fading i.e., coherence time of channel is equal to one data transmission block duration. It is clear that our modified technique outperforms the original technique, which is clearly in coherence with the discussion of section (3.3.1).

In the Figure 3.5, the plot shows the comparison of performance between the systems when our modified DF-DD is applied and when [40] is applied. The channel considered is slow fading i.e., coherence time of channel is equal to two data transmission blocks duration. It can clearly be seen that, our modified technique performs better than original technique,

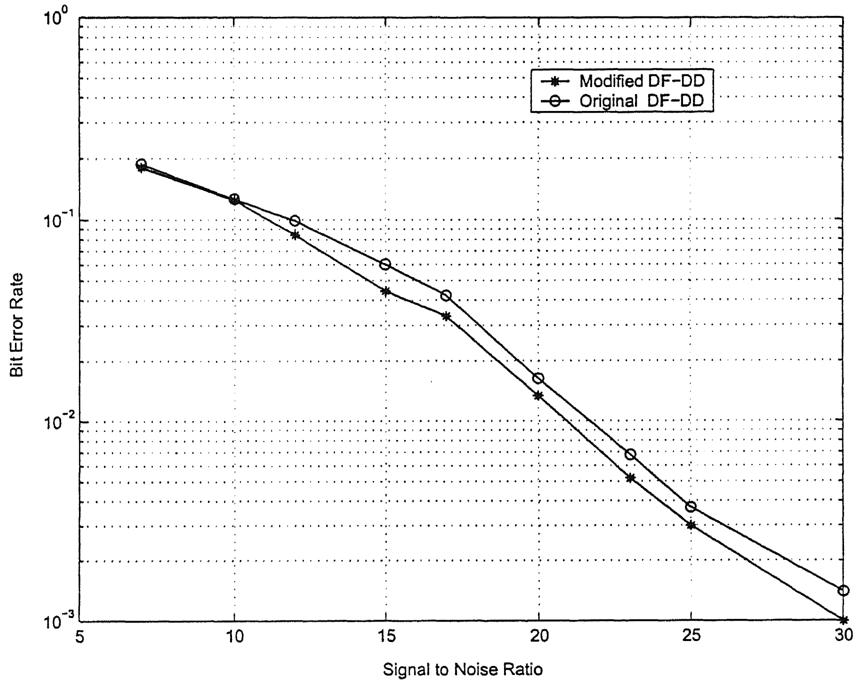

Figure 3.5: Comparison of system performance when modified DF-DD is applied and when original DF-DD [40] is applied in slow fading channel

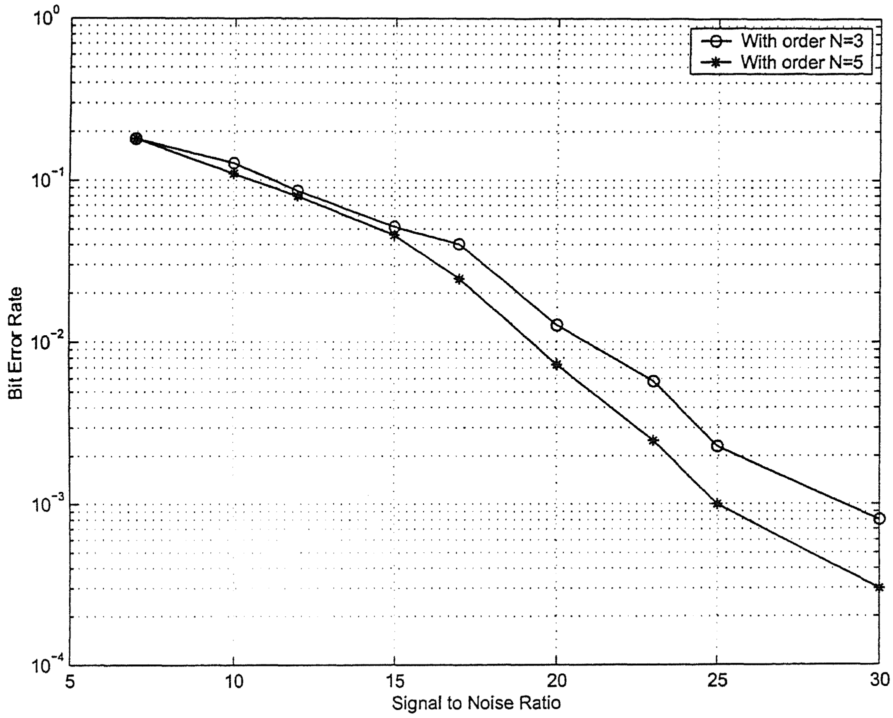

Figure 3.6: Plot showing the comparison of performance between DOSTM-DD systems employing LMS adaptive algorithms with 2 different orders $N = 3$ and $N = 5$ in slowly fading channel.

which is clearly in coherence with the discussion of section (3.3.1).

The Figure 3.6 we show the performance of LMS used for adaption process for DF-DD, instead of RLS. The channel considered for this plot is slow fading channel. we can observe that the proposed LMS based scheme is as good as the RLS based scheme of [40], if proper initialization of RLS algorithm is considered. It is to be noted here that LMS significantly reduces the complexity of the system. The plot is shown for both order, $N = 3$ and $N = 5$.

The Figure 3.7 shows performance of LMS based adaption scheme in fast fading channels. It can be observed (from earlier plots), that the proposed LMS based scheme is performing as good as the RLS based scheme of [40] with properly chosen initialization better.

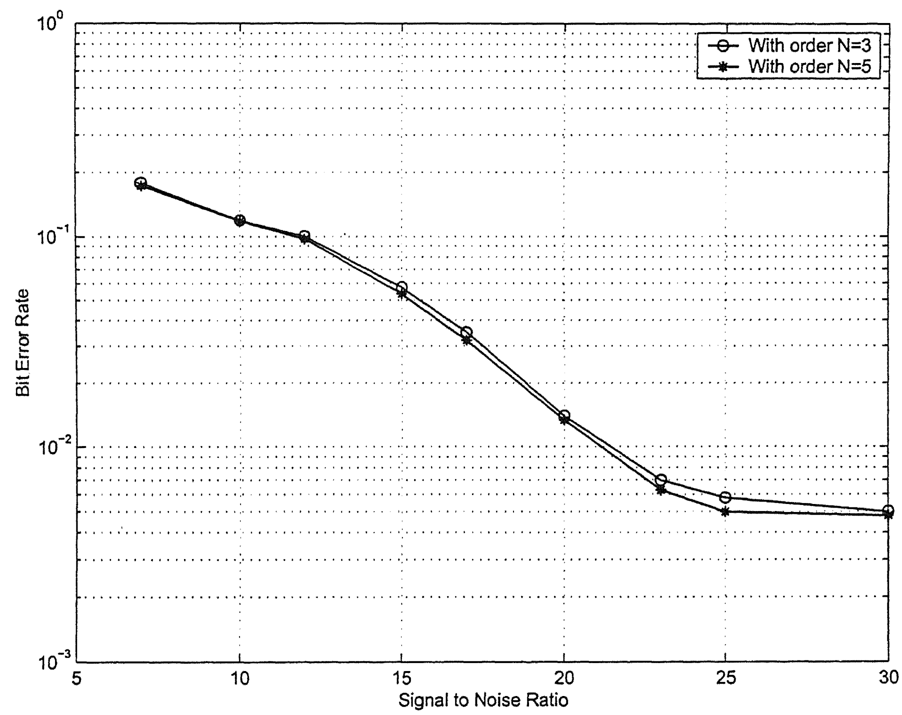

Figure 3.7: Plot showing the comparison of performance between DOSTM-DD systems employing LMS adaptive algorithms with 2 different orders $N = 3$ and $N = 5$ in fastfading channel.

Chapter 4

OFDM DF-DD for Frequency Selective channels

Orthogonal frequency-division multiplexing (OFDM) has recently received increased attention due to its capability of supporting high-data-rate communication in frequency selective fading environments which cause inter-symbol interference (ISI). The basic idea of serial-to-parallel converting of data and transmitting over orthogonal frequency channels has started long back. The [44] and [45] are some of the pioneering papers of this direction.

Instead of using a complicated equalizer as in the conventional single carrier systems, the ISI in OFDM can be eliminated by adding a guard interval which significantly simplifies the receiver structure. However, to reap all the benefits offered by OFDM, it has to be combined with proper coding technique. Therefore, coding becomes an inseparable part in most OFDM applications and a considerable amount of research has focused on optimum encoder, decoder, and interleaver design for information transmission via OFDM over fading environments. Space time code combined with OFDM has attracted much researchers in the past few years. For the growing need for high quality and high reliable links, apparently OFDM clubbed with high diversity space time codes is better option to go.

In this chapter we first detail about the OFDM technique and then see the effect of combining our DF-DD scheme with OFDM to make it work for more adverse, frequency selective environments. It is to be noted, that in all the discussions made till now, we have

assumed that channel is frquency flat faded (i.e., channel doesn't have memory), which is untrue for the practical case. If another new layer OFDM is added to the existing system, we can make the system work well even in the frequency selective fading channels(i.e., the channels with memory).

4.1 OFDM system model

The basic principle of OFDM is to split a high-rate datastream into a number of lower rate streams that are transmitted simultaneously over a number of subcarriers. Because the symbol duration increases for the lower rate parallel subcarriers, the relative amount of dispersion in time caused by multipath delay spread is decreased. Intersymbol interference is eliminated almost completely by introducing a guard time in every OFDM symbol. In the guard time, the OFDM symbol is cyclically extended to avoid intercarrier interference. The basic block diagram showing the OFDM modulation scheme is shown in the Figure 4.1

As shown in the Figure 4.1, an OFDM signal consists of a sum of subcarriers that are modulated by using Phase Shift Keying (PSK). In the literature [46] complex baseband notation is used to express the OFDM signal

$$s(t) = \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}} d_{i+N_s/2} e^{(j2\pi\frac{i}{T}(t-ts))}, \; t_s \le t \le t_s + T \tag{4.1}$$

Where N_s is the number of subcarriers, T is the OFDM symbol duration, and f_c is the carrier frequency and d_i are the complex PSK symbols.

The real advantage of the OFDM comes in the ease of implementation of the above signal. $s(t)$ in (4.1) is in fact nothing more than the inverse Fourier transform (IDFT) which can be easily implemented using FFT algorithm.

One of the most important reasons to do OFDM is the efficient way it deals with multipath delay spread. By dividing the input datastream in N_s subcarriers, the symbol duration is made N_s times smaller, which also reduces the relative multipath delay spread. To eliminate the inter symbol interference completely, a guard time is introduced for each OFDM symbol. The guard time is choosen larger than the expected delay spread, such that the multipath components from one symbol cannot interfere with the next symbol. The guard time could

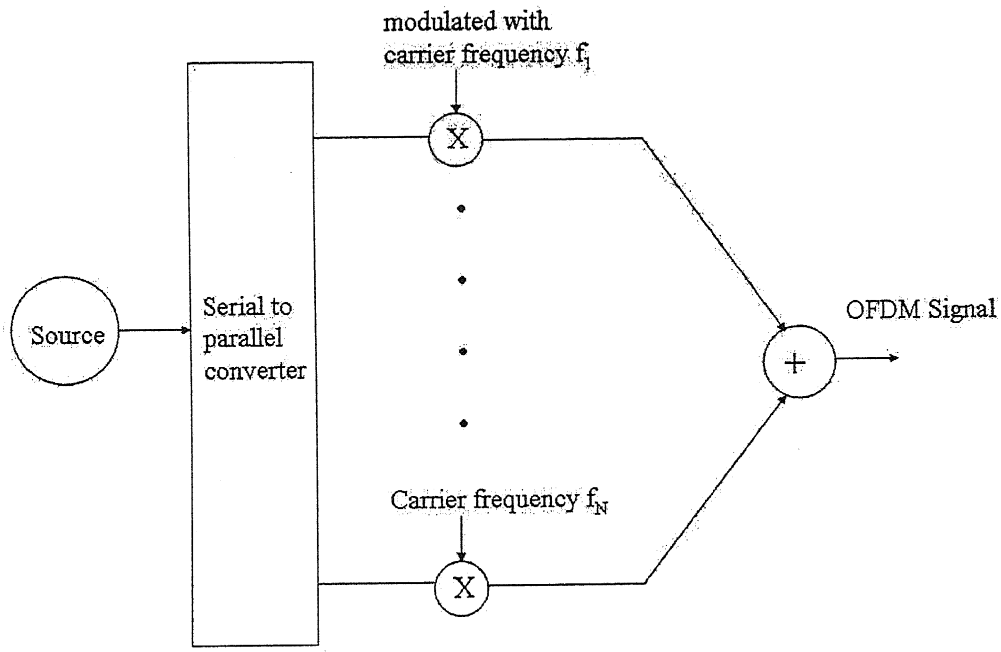

Figure 4.1: OFDM modulator block diagram

consist of no signal at all(which is called as Zero Padded OFDM). In some cases the OFDM symbol is extended into the guard time (which is called as Cyclic Prefix OFDM). In our case we choose the ZP-OFDM (Zero Padded OFDM) owing to it's simplicity and ease of implementation.

4.2 Channel Modelling

The channel considered for the system in this chapter is frequency selective fading channel. It is modelled from modified Jakes' algorithm for Rayleigh frequency non-flat fading channels. Modelling of a frequency non-flat fading channel is more complicated compared to the that of flat fading channel owing to channel's memory. The channel is said to be of order L if it has the memory of previous L symbols put on it and the channel coefficients would be $L+1$ in number. Modified Jakes' model for Rayleigh channels is widely in use to simulate the frequency selective fading channel, owing to it's simplicity and efficiency. Here, in our case we require multiple uncorrelated channels, which is not straight forward to simulate. Initial channel model proposed by Jakes [47] in 1974 is deterministic model for only one channel, This was extended for multiple channels using almost same technique by Dent et.,all in [48] in 1993. In this the authors reformulated the Jakes' Model with slightly different arrival angles of the incident multipath rays and apply orthogonal weighting functions (Walsh-Hadamard codewords) on the oscillators in the model to produce uncorrelated outputs. Again in 2000, this problem is revisited by yingbo Li and Y. L. Guan in [49], in which the authors suggested a much simpler structure that do not require the use of orthogonal weighting functions. The technique explained in [49] is to generate independent fading signals by using the same set of sinusoidal generators, but with a suitable set of phase shifts on each of them. The key advantages of this model is, its greatly reduced executing time and capability for simultaneous generation of multiple uncorrelated fading signals.

4.2.1 Structure of channel Simulator

Here we shall discuss the complete structure of the simulator, the algorithm used for developing the model.

In this model, we generate fading signal as a discrete sum of in-phase and quadrature phase sinusoidal signals with different Doppler frequencies and appropriate phase shifts. we adopt symmetrical values for the Doppler shift components (w_n) and initial phase shift (θ_{nj}) as in [48]. The in-phase and quadrature-phase components of the jth path fading waveform are, respectively:

$$X_{cj}(t) = 2\sum_{n=1}^{N_o} cos(\theta_{nj})cos(w_n t + \pi j/2) \tag{4.2}$$

and

$$X_{cj}(t) = 2\sum_{n=1}^{N_o} sin(\theta_{nj})cos(w_n t + \pi j/2) \tag{4.3}$$

Where,

$$\begin{aligned}
\theta_{nj} &= \frac{\pi n j}{N_o},\ j = 1, 2, ..., N_o - 1. \\
w_m &= 1\pi V/\lambda;\ V = \text{Vehicle speed}; \\
\lambda &= \text{Carrier wavelength}; \\
w_n &= w_m \cos\frac{2\pi(n-0.5)}{N}; \\
N &= 4(N_o + 1)\ N_o = \text{Number of oscillators.}
\end{aligned}$$

The selection for θ_{nj}, is to ensure that the in-phase and quadrature-phase components have equal average power and are uncorrelated. The additional phase $\pi J/2$ is imposed to cancel out the cross correlation among quadrature components of different paths. The choice of N is to give same number of oscillators (No) as the classical Jakes' model.

The structure diagram is shown in the Figure 4.2 depicts the structure for the jth path of the model

4.2.2 Validating the simulated channel model

In this section, we validate the channel model explained in the previous section, by plotting the autocorrelation of the fading coefficients generated by simulated model. We also show the plot of theoretical autocorrelation function plot of Rayleigh non-flat fading channel coefficients for comparison.

In the Figure 4.3 we have plotted the theoretical autocorrelation of Rayleigh non-flat faded channel coefficients. The mathematical expression for this autocorrelation is given by

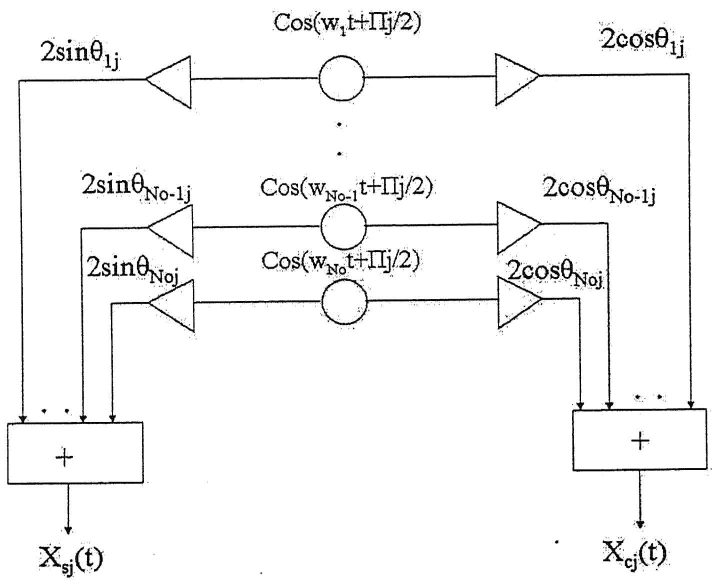

Figure 4.2: Flow chart showing the structure of the simulator for generating non-flat fading coefficients from modified Jakes' model

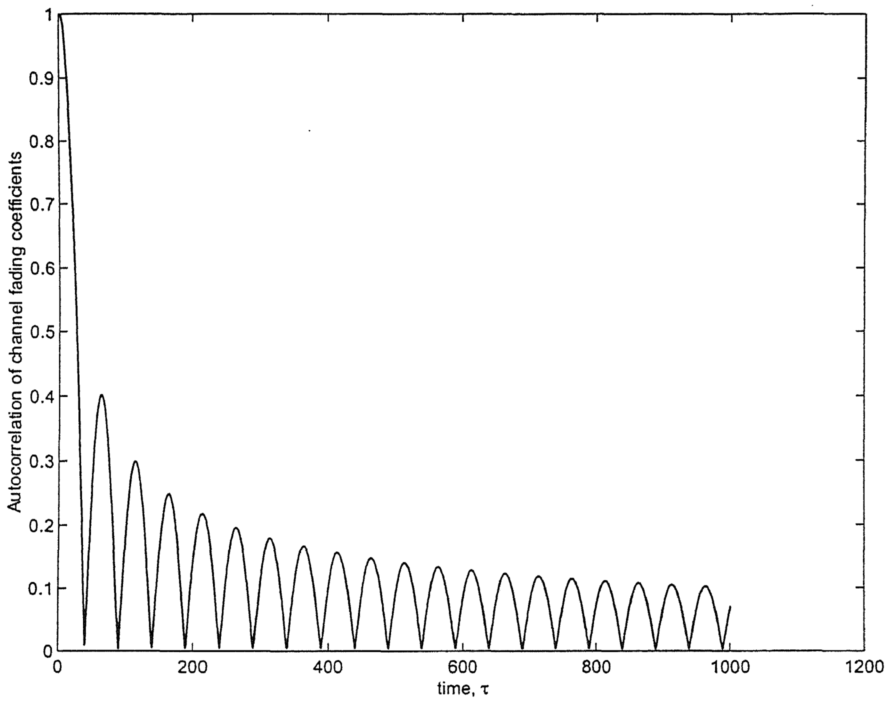

Figure 4.3: Plot showing the theoretical autocorrelation function of the Rayleigh non-flat faded channel coefficients. The expression for autocorrelation is $r_h(t) = J_0(2\pi f_d T_s t)$.

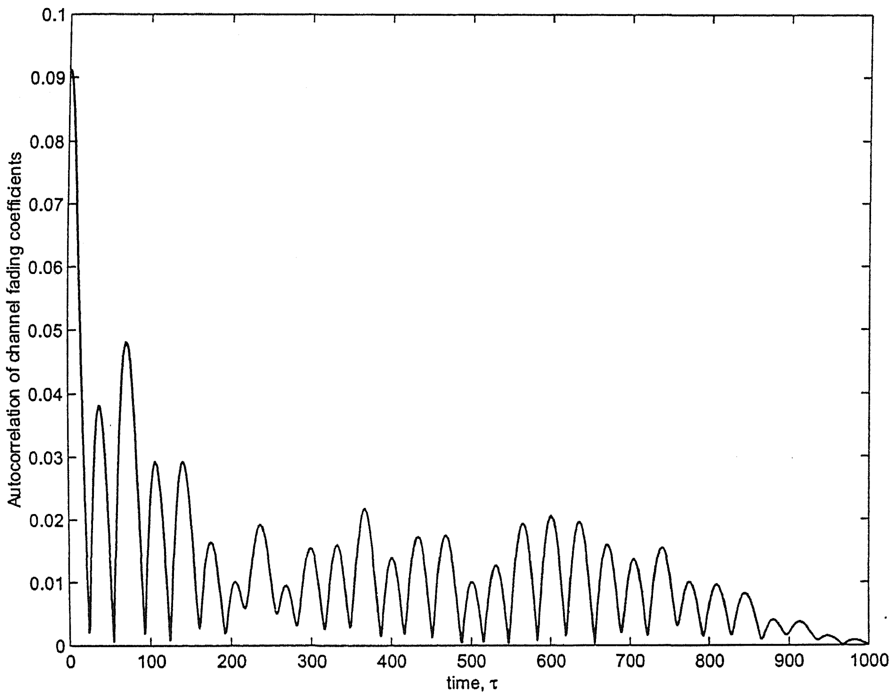

Figure 4.4: Plot showing the autocorrelation of simulated channel

$r_h(t) = J_0(2\pi f_d T_s t)$, where J_0 is the bessel function of first kind and order zero, f_d is the maximum doppler frequency shift, T_s is the the bit duration.

In the Figure 4.4, we have plotted the autocorrelation of the simulated channel from the model of explained in the previous section. It can be observed that this is in agreement with the autocorrelation of theoretical Rayleigh non-flat fading channel.

4.3 OFDM with Decision Feedback-Differential Detection

As said earlier, the DF-DD for DOSTM system has the limitation of being applicable only to the flat fading channels. In this work, we propose to include the new layer OFDM into the DF-DD for DOSTM system to make it combat the non-flat(frequency selective fading)

channels. The basic principle involved is, OFDM turns the frequency selective channel into a set (N) of flat fading channels, over each of which our previous DF-DD scheme is applied. Inclusion of OFDM in differential techniques is not staright forward. The existing blocks of the system and blocks of OFDM layer are interleaved, unlike the other systems which employ OFDM in the last layer to combat frequency selective fading channels.

The Figure 4.5 shows the description complete block diagram of OFDM DF-DD for DOSTBC system.

Description: We first serial to parallel convert the APSK modulated symbol stream, into $2p$ parallel channels, where p is the number of subcarriers used for OFDM modulation. Then the 2 symbols, one from each set of p symbols data from $2p$ parallel streams are given to the DSTBC block. There are p such DSTBC blocks, each of which employs the encoder as explained in the earlier chapters. Each block gives the 2 symbol for each symbol duration T. Then the outputs of all the blocks are taken and rearranged to make them ready for the OFDM modulation using IFFT. Finally Zero padding is done, since we are employing Zero Padding OFDM (ZP-OFDM).
The OFDM symbol thus generated is put on the channel. Channel is simulated according to the modified jakes' model for non-flat fading, which is described in detail in the earlier section of this chapter. The decoding of the OFDM symbols and subsequent DF-DD is done in the exact reverse order as explained in the block diagram. At the receiver, first zero padding is removed. Then FFT is applied on each of the blocks separately, after which rearranging of the symbols is done before actual DF-DD is applied for decoding the symbols. Finally parallel to serial convertion is done for stream of detected APSK symbols.

4.4 Simulation Results

In this section we show the simulation results for the performance evaluation of the above described OFDM DF-DD DOSTBC system. Since the channel conditions for the earlier discussed system (DF-DD for DOSTBC) and for this system are not same, we cannot compare the performance of these two systems. In order to evaluate the performance of this OFDM based system, we have considered a totally different system which works on similar channel

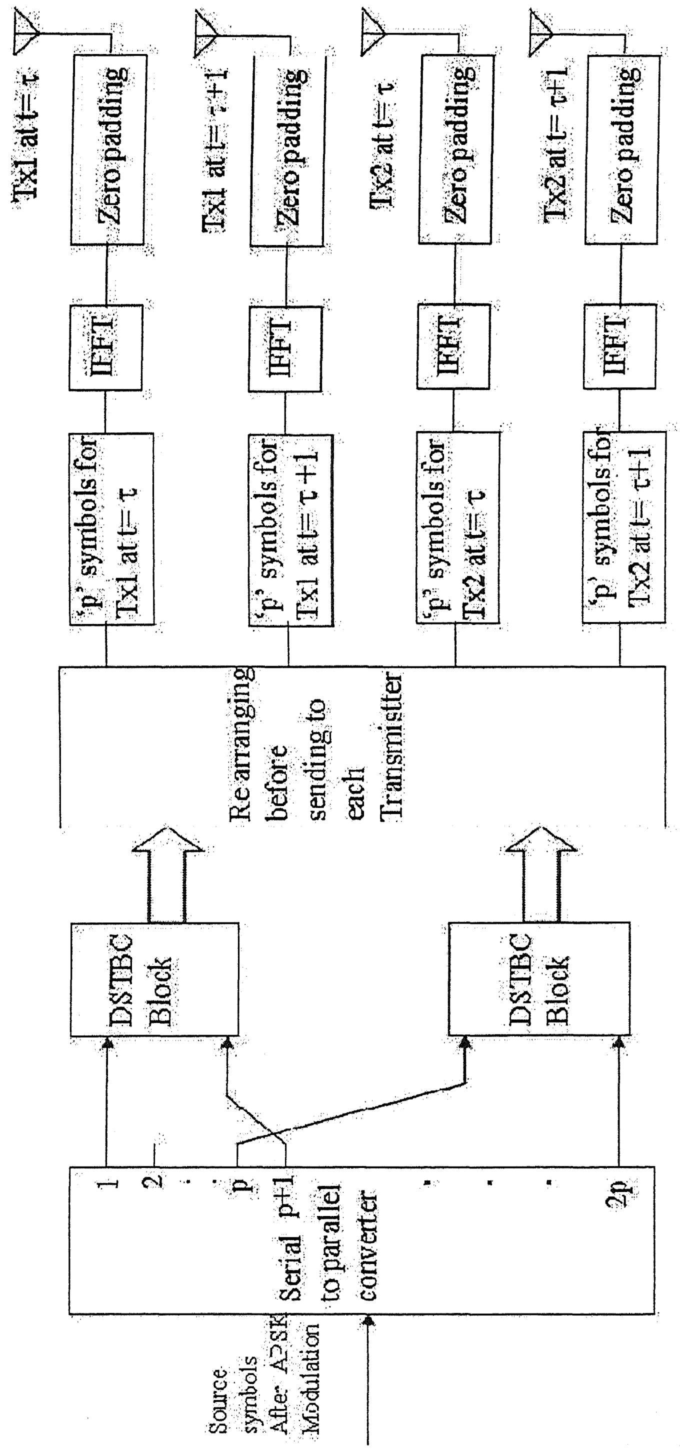

Figure 4.5: The block diagram showing the complete system model of OFDM based DF-DD for DOSTBC system

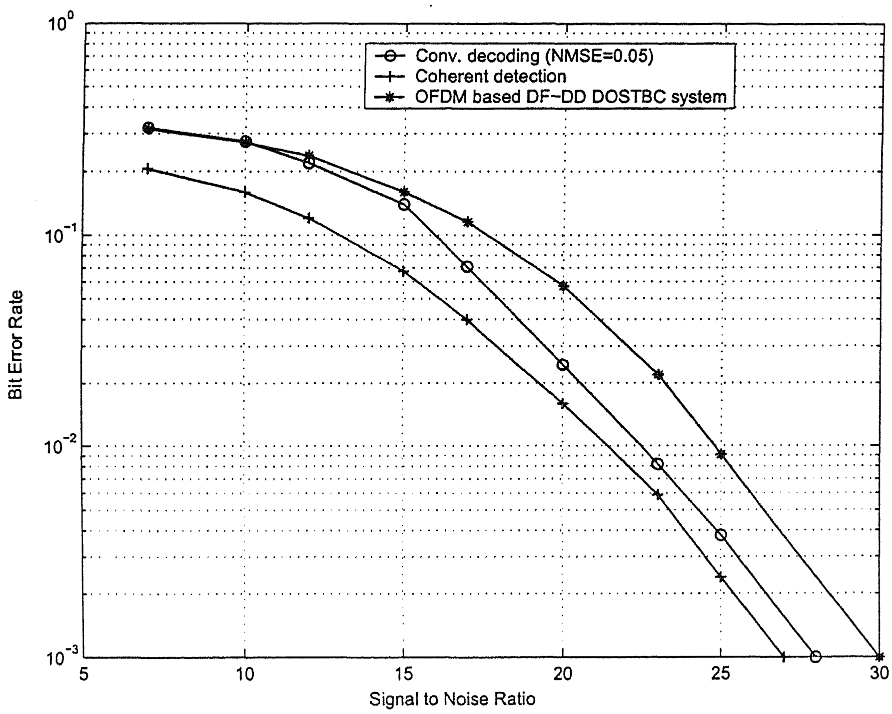

Figure 4.6: Performance of the proposed OFDM based system compared to conventional STBC decoding with channel estimation error NMSE=0.05. The plot of coherent detection, with perfect channel estimation is also shown in the figure.

conditions. The system under consideration is from the work in [18]. This is a conventional STBC system employing OFDM to combat the frequency selective fading channels. The decoding is done after channel estimation by complexity intensive sub-space based semiblind method. This estimation method requires less number of pilot symbols when compared to the other pilot based methods available.

In the Figure 4.6 we have plotted the performance curve for the three systems. First is the proposed OFDM based DF-DD DOSTBC system, second is conventional STBC decoding with NMSE of channel estimation as 0.05, and coherent detection of OSTBC with perfect channel estimation. It is evident that the proposed system performance is not better than the conventional system, but it is to be noted that differential schemes have a theoretical

performance loss of about 3dB when compared to their non-differential counterparts. The conventional decoding system considered has some disadvantages compared to our system they can be listed as:

1) It requires channel to be constant for the whole duration of time, for channel estimation and subsequent symbol detection. This could be as high as the duration of 20 blocks of data transmission.

2) The implementation complexity is quite cumbersome compared to our proposed system, with subspace based channel estimation. The estimation technique have some complexity intensive mathematical operations such as SVD(Singular Value Decomposition), which takes considerable amount of time and processor resources.

3) It uses the pilot symbols for channel estimation. This definitely shares the bandwidth allocated for data transmission, however small it may be.

It can also be noted from the graph that the difference between the coherent detection and proposed system is around 3dB. This is completely in agrement with what we have discussed in earlier chapters about the performance loss in general for differential techniques.

Chapter 5

Conclusion

5.1 Conclusion

In this thesis, we have studied the APSK signalling scheme applied to Differential Orthogonal Space Time Block codes and also a particular decoding technique based on Decision Feedback adaptive filtering. We have suggested a modification to the existing decoding technique of Decision Feedback Differential Detection for Orthogonal Space Time Block codes (DF-DD DOSTBC). This modification is supported by a theoretical proof. The simulation results showing the performance of the modified system and existing system also corroborate the claim.

Then we have studied the effect of replacing relatively complex, Recursive Least Square(RLS) algorithm in the adaptive filtering of DF-DD for DOSTBC with much simpler and easy to implement Least Mean Square (LMS) algorithm. Then the performance of both the systems are evaluated in different channel conditions and with different initialization parameters.

The DF-DD for DOSTBC scheme available so far is applicable only for flat-fading, memoryless channels (frequency non-selective). We have included a new layer in the system with OFDM modulation and enabled the whole system to work for more practical channels, i.e., frequency selective fading channels. The channel used for simulations is modelled by modified Jakes' model for Rayleigh frequency selective fading channels.

Regarding the future scope of work, the performance evaluation of the proposed systems

using analytical techniques can be attempted. Similarly the theoretical expressions for error probability as a function of Signal to Noise Ratio (SNR) is can be derived.

References

[1] R. G. Gallager, *"Information Theory and Reliable Communication,"* J wiley and sons, New York, 1968.

[2] I.E. Telatar,"Capacity of multi-antenna Gaussian channels," *Euro. Trans. Telecommun.*, vol.10, pp.585-595, November.

[3] G. J. Foschini and M. J. Gans, "On Limits of Wireless Communications in a Fading Environment" *Wireless personal communications* , vol.6, no.3, pp. 311-335, March 1998.

[4] P. W. Wolniansky, G. J. Foschini, G. D. Golden, and R. A. Valenzuela, "V-BLAST- an architecture for realizing very high data rates over the rich-scattering wireless channel" *Proc. ISSSE-98, september 1998.*

[5] G. J. Foschini, "Layered space-time architecture for wireless communication in a fading environment when using multi-element antennas," *Bell Labs Technical journal,*Vol. 1, no. 2, pp. 41-59, Autumn 1996.

[6] S. M. Alamouti, "A Simple transmit diversity technique for wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, no. 8, October 1998.

[7] A. Paulraj, R. Nabar and D. Gore, Simpson, *Introduction To Space-Time Wireless Communications.*Cambridge University press 2003.

[8] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communication: performance analysis and code construction," *IEEE trans. Inform. Theory*, pp.744-765, March 1998.

[9] P. Stoica and G. Ganesan, "Trained space-time block decoding for flat fading channels with frequency offsets," *Wireless Personal communications*, Vol. 27, no. 3, pp. 235-245, 2003

[10] Z. Liu, X. Ma and G. B. Giannakis, "Space-time coding and kalman filtering for time selective fading channels," *IEEE Trans. Commun.*, Vol. 50, no. 2, February 2002.

[11] Z. Liu, X. Ma, G. B. Giannakis, and B. L. Hughes, "Double differential space-time block coding for time selective fading channels," *IEEE Trans. Commun.*, Vol. 49, no. 9, September 2001.

[12] Vahid Tarokh and Hamid Jafarkhani "A differential detection scheme for transmit diversity," *IEEE J. Select. Areas Commun.*, VOL. 18, NO. 7, JULY 2000

[13] A. Wittneben, Base station modulation diversity for digital SIMULCAST, *Proc. IEEE VTC*, pp. 848853, May 1991.

[14] N. Seshadri and J. H. Winters, Two signaling schemes for improving the error performance of frequency-division-duplex (FDD) transmission systems using transmitter antenna diversity, *Int. J.Wireless Inform. Networks*, vol. 1, no. 1, 1994.

[15] J.-C. Guey, M. P. Fitz, M. R. Bell, and W.-Y. Kuo, Signal design for transmitter diversity wireless communication systems over Rayleigh fading channels, *Proc. IEEE VTC*, pp. 136140, 1996.

[16] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, Space-time block codes from orthogonal designs, *IEEE Trans. Inform. Theory*, vol. 45, pp. 14561467, July 1999.

[17] V. Tarokh, H. Jafarkhani, and A. R. Calderbank, Space-time block codes for high data rate wireless communication: Performance results, *IEEE J. Select. Areas Commun.*, vol. 17, pp. 451460, Mar. 1999.

[18] Yonghong Zeng and Tung Sang Ng, "Subspace-based semi-blind channel estimation for STC-OFDM", *Intern. Conf. on Communi., ICC*2004.

[19] Z. Liu, G. B. Giannakis, S. Barbarossa, and A. Scaglione, Transmit antennae space-time block coding for generalized OFDM in the presence of unknown multipath, *IEEE J. Select. Areas Commun.*, vol. 19, no. 7, pp. 13521364, 2001.

[20] A. Scaglione, G. B. Giannakis, and S. Barbarossa, Redundant filterbank precoders and equalizers-part ii: blind channel estimation, synchronization and direct equalization, *IEEE Trans. Signal Processing,* vol. 47, no. 7, pp. 20072022, 1999.

[21] E. Moulines, P. Duhamel, J. F. Cardoso, and S. Mayrargue, Subspace methods for the blind identification of multichannel FIR filters, *IEEE Trans. Signal Processing,* vol. 43, no. 2, pp. 516525, 1995.

[22] Y. H. Zeng and T. S. Ng, A semi-blind channel estimation method for multi-user multi-antenna OFDM systems, *IEEE Trans. Signal Processing,* vol. 52, no. 5, 2004.

[23] Z. Ding and D.B. Ward, "Semi-blind equalization for space time block codes and its ambiguity analysis", *Proc. IEEE Int. Conf. Communicat. (ICC-04),* vol. 4, pp. 2357-2361, Paris, June 2004.

[24] B. M. Hochwald and T. L. Marzetta, Unitary spacetime modulation for multiple-antenna communications in rayleigh flat fading, *IEEE Trans. Inform. Theory,* vol. 46, pp. 543564, 2000.

[25] Xiang-Gen Xia, Differentially en/decoded orthogonal space-time block codes with APSK signals, *IEEE Commun. Lett.*, vol. 6, pp. 150152, Apr. 2002.

[26] B. M. Hochwald and W. Sweldens, Differential unitary space-time modulation, *IEEE Trans. Commun.*, vol. 48, pp. 20412052, Dec. 2000.

[27] W. T. Webb , Hanzo L., Steele R., "Bandwidth Efficient QAM Schemes for Rayleigh Fading Channels," *IEE Proceedings-1,* vol. 138, No. 3, pp. 169-175. Jun. 1991.

[28] Y.C. Chow, A.R. Nix, and J. P. McGeehan "Analysis of 16-APSK modulation in AWGN and Rayleigh fading channel" *Electronics Letters* Volume 28, Issue 17, pp. 1608 - 1610 Aug. 1992

[29] Rohling, M.; May, T.; Bruninghaus, K.; Grunheid, R. "Broad-band OFDM radio transmission for multimedia applications," *Proceedings of the IEEE* Volume: 87, Issue: 10, pp. 1778-1789, Oct 1999

[30] Aijun Song, Genyuan Wang, Weifeng Su, and Xiang-Gen Xia, "Unitary space-time codes from Alamouti's scheme with APSK signals," *Wireless Communications, IEEE Transactions on* Volume: 3, Issue: 6, pp. 2374- 2384, Nov. 2004

[31] F. Adachi, M. Sawahashi, "Decision Feedback Differential Detection of Differentially Encoded 16APSK Signals," *IEEE Trans. Commun.*, Vol. 44, No. 4, pp. 416-418, Apr. 1996,

[32] B. M. Hochwald and W. Sweldens, "Differential unitary space-time modulation," , *IEEE Trans. Commun.*, Vol. 48, pp. 2041-2052, Dec. 2000.

[33] I. Korn, Error floors in the satellite and land mobile channels, *IEEE Trans. Commun.*, vol. 39, pp. 833837, June 1991.

[34] R. Schober, W. H. Gerstacker, and J. B. Huber, Decision-feeback differential detection of MDPSK for flat Rayleigh fading channels, *IEEE Trans. Commun.*, vol. 47, pp. 10251035, July 1999.

[35] R. Schober and W. H. Gerstacker, Decision-feeback differential detection based on linear prediction for MDPSK signals transmitted over Ricean fading channels, *IEEE J. Select. Areas Commun.*, vol. 18, pp. 391402, Mar. 2000.

[36] F. Adachi and M. Sawahashi, Decision feeback differential detection of 16-DAPSK signals, *Electron. Lett.*, vol. 29, no. 16, pp. 14551457, Aug. 1993.

[37] R.-Y.Wei and M. C. Lin, Modified decision feeback differential detection for differentially encoded 16-APSK signals, *Electron. Lett.*, vol. 34, no. 4, pp. 336337, Feb. 1998.

[38] R. Schober,W.H. Gerstacker, and J. B. Huber, Decision-feeback differential detection based on linear prediction for 16 DAPSK signals transmitted over flat Ricean-fading channels, *IEEE Trans. Commun.*, vol. 49, pp. 13391342, Aug. 2001.

[39] Y. Liu and X. Wang, Multiple-symbol decision-feedback space-time differential decoding in fading channels, *EURASIP J. Appl. Signal Process.*, vol. 3, pp. 297304, Mar. 2002.

[40] Aijun Song and Xiang-Gen Xia, "Decision Feedback Differential Detection for Differential Orthogonal Space-Time Modulation With APSK Signals Over Flat-Fading Channels," *IEEE Transactions on wireless communications*, vol. 3, pp.1873-1878, Nov. 2004.

[41] S. Haykin, *Adaptive Filter Theory*, 3rd ed. Englewood Cliffs, NJ: Prentice- Hall, 1996.

[42] M. Patzol and F.Laue "Statistical properties of Jakes' fading channel simulator" *48th IEEE Vehicular Technology Conference*, 1998. VTC 98. Volume 2, pp. 712 - 718, 18-21 May 1998

[43] M. F. Pop and N. C. Beaulieu, "Statistical investigation of sum-of-sinusoids fading channel simulators" *Global Telecommunications Conference, 1999*, GLOBECOM '99 Volume 1A, pp. 419 - 426, 1999.

[44] S. B. Weinstein and P. M. Ebert, "Data transmission by frequency-division mul- tiplexing using the discrete Fourier transform," *IEEE Trans. Commun. Technol.*, vol. COM-19, pp. 628-634, Oct. 1971.

[45] J. A. C. Bingham, "Multicarrier modulation for data transmission: An idea whose time has come," *IEEE Commun. Mag.*, vol. 28, pp. 5-14, May 1991.

[46] Richard Van Nee and Ramjee Prasad. *OFDM For Wireless Multimedia Communications*, Artech House Publishers, Boston-London, 2000.

[47] Jakes W. C. *Microwave Mobile Communications*, John Wiley and Sons, Newyork, 1974.

[48] P. Dent, G.E. Bottomley, and T. Croft "Jakes Fading Model Revisited", *Electronics Letters*, Vo1.29, No.13,24 June 1993, pp.1162-1163.

[49] Yingobo Li and Y. L. Guan, "Modified Jakes' Model for Simulating Multiple Uncorrelated Fading Waveforms" *IEEE 51st Vehicular Technology Conference Proceedings, 2000, Tokyo*. Volume 3, pp. 1819 - 1822.

[50] John G. Proakis, *Digital Communications*, 3rd ed. McG RAW-HILL, Electrical Engineering series, 1995.